0041

Development of a portable AC/DC welding power supply module.

Bethlehem Steel Corporation

FOREWORD

The purpose of this report is to present the results of one of the research and development programs which was initiated by the members of the Ship Production Committee of The Society of Naval Architects and Marine *Engineers* and financed largely by government funds through a cost sharing contract between the U.S. Maritime Administration and Bethlehem Steel Corporation. The effort of this project was directed to the development of improved methods and hardware applicable to shipyard welding in the U.S. shipyards.

Mr. W. C. Brayton, Bethlehem Steel Corporation was the Program Manager, Mr. G. E. Iayer, Celesco Industries Inc. directed the development work at the San Diego, California plant. Grateful acknowledgement is made for the outstanding contribution of Messrs. J. Thommes, H. White, axial G. Silke of the Celesco organization.

Special acknowledgement is made to the members of Welding Panel SP-7 of the SNAME Ship Production Committee who served as technical advisors in the preparation of inquiries and evaluation of sub-contract proposals.

DEVELOPMENT OF A PORTABLE AC/DC POWER SUPPLY MODULE

PROJECT SP-1-2 (203.1)

A TECHNICAL REPORT SUBMITTED TO THE

BETHLEHEM STEEL CORPORATION

celesco

industries inc.

ENVIRONMENTAL AND INDUSTRIAL PRODUCTS DIVISION

7520 Convoy Court • San Diego • California 92111

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ABSTRACT

This report deals with lightweight power conversion for application and portability in the welding industry. It is a final report on a research contract and it includes all the circuits investigated and their potentials. The report also includes a detail description of the final lightweight circuit used. The supply has a maximum output of 250 amps at 30 volts and either voltage or current could be regulated by a feedback control system. Ninety percent variability is possible with 85 percent efficiency at full load. The unit weighs 98 pounds and is convection cooled only. Input power is 440 VAC 60 Hz 30.

A) EXECUTIVE SUMMARY PROGRAM HISTORY

January 1973	System approach under investigation used transistor switches in bridge configuration. Welding was accomplished with this system; however, problems were encountered with transistor breakdown and overheating of switches. Failure of input rectifiers was also experienced. All waveforms displayed large amounts of noise. Efficiency was calculated at 65%.
February 1-15 1973	A number of failures of both switches and rectifiers were encountered. Input bridge rectifier types were changed to 1N2287. A new transformer was wound with an increased number of turns in the secondary to provide a higher voltage under load.
February 16-28 1973	Preliminary investigation was begun on an SCR switching circuit. Also, parallel tuned LC circuits were inserted in the primary side of the trarsistor switch transformer to correct the switching waveform and reduce harmonics.
March 1973	A short to the core of the power transformer caused destruction of the core. At this time, a smaller core was obtained and the winding configuration recalculated. The SCR swithes were run at low voltage using capacitor commutation. Problems were encountered with lock-up. A series switch configuration was suggested to reduce the voltage drop on each transistor. This circuit was breadboarded and iests started. The SCR evaluation was deferred.
April 1973	Work was continued on the serial transistor switch approach. Problems were found in turning off both series transistors at exactly the Same time. Minor differences caused full voltage to appear across one unit, resulting in its destruction. Transistor type used was RCA 2N6251. Increased transistor drive with circuits to remove stored base charge more rapidly on rum off failed to make this system operational. DC biasing of the transistor operating at a higher level was not feasible due to the large amount of power dissipated in the bias circuit. Delco engineers suggested DTS 804 for this application. Circuit was rebuilt using this single transistor in place of the series pair. A parallel investigation was begun on using SCR switches.

May 1973	The transistor switch configuration was operated at efficiencies in excess of 88%; however, voltage and current transients still caused switch destruction. At this time, it was decided that the remainder of the month would be devoted to the SCR intestigation. A switch commutated system was arrived at which shows considerable promise in this application. Spec sheets *were* received on a new Motorola Beam-fired SCR with ratings well in excess of the requirements. Samples were requested.
June 1973	The transistor switch in bridge configuration was selected as the best means of achieving success within budget and schedule. Concentrated effort was applied to this circuit, using DTS 804's. AS a result of careful timing and wave-shaping, switch dissipation was reduced to a minimum. The system was run for five hours at 100 amperes output without significant heating of the bridge. Input voltage was increased incrementally, with waveform observation and correction being made at each level. The circuit was operated at full input voltage. Load transient protection and current limiting circuitry were incorporated in preparation for tests with an arc.
July 1973	Saturable reactor current limiters were developed for current transient limitation. The reactors made a con siderable weight contribution of about 20 pounds; however they were very successful at limiting current transients. As a result the output current could be increased to 250 amp at full voltage without destruction of the semiconductors.
August 1973	The complete breadboard was tested under welding conditions at full voltage and current; however, at this higher power level, excessive heating was observed. The breadboard was folded into a mechanical package and tested under welding conditions (SMAW). Problems at rapid turn-on of this input line caused destruction of semiconductor switches.
September 1973	A Zero crossover SCR switch was incorporated into the input rectifier to reduce the secondary peaks encountered in rapid turn-on. Moderate success was achieved.
October 1973	Additional bypass capacitors were added to reduce further switch heating and voltage transients.

November 1973	AC welding was achieved with limited success due to the high frequency of transformation. voltage process was added to the unit. Mig welding was successfully performed.
December 12 1973	Successful demonstration to Marad Commission and visitors in Bethlehem Steel Sparrows Point Shipyard.

B) CONCLUSIONS

From a welding point of view, the final unit fulfilled a substantial portion of the desired objectives.

From an economic and reliability point of view, the transistor design is not optimum. It is felt SCR's are a better device for this application, providing they can be controlled.

Recommendations for future study:

1) Investigate the use of 220V instead of 440V.

2) In the light of techniques used in this project, re-investigate the use of SCR's.

3) Investigate the conversion of inversion frequency back to 60 Hz for AC welding.

4) Investigate a 120V single phase mini supply.

I Objectives

A) Goals

The following is a statement of the desired performance of a light-weight power supply for welding applications.

The basic power supply module shall be capable of supplying a min-mum of 250 amperes of welding current suitable for AC shielded metal arc welding within that power requirement.

The basic module shall be adaptable to provide a minimum DC welding current of 250 amperes of parallel connected with other modules to provide up to 1000 amperes DC for shielded metal arc and submerged arc welding.

The primary module or combinations of the primary module shall be designed to support shipyard production welding. Production welding includes the following types of operation:

1. Automatic submerged arc welding of plate butts in downhand position.
2. Semi-automatic sub-arc and/or MIG welding of butt weld and large fillets which are not accessible to machine driven equipment.
3. Manual stick electrode welding using large diameter electrodes for downhand fillets and short butts. Gravity welding is also in this category.
4. Electro-slag and/or electro-gas welding of vertical butts in shell and bulkheads.
5. Consumable nozzle welding of vertical butts in stiffeners and girder
6. All position MIG welding with small diameter electrode.
7. All position stick electrode welding.
8. Small amounts of special steel and non-ferrous welding both MIG and manual.
9. Arc gouging second sides of butt welds and correcting defective wel
10. Plasma-arc operations.

Primary and intra-modular commotions shall be of a plug-in type where practical so as to minimize the use of a skilled technician for maintenan

Primary power sourer shall be 440-480 volt three phase 60 cycle AC. Consideration may be given to higher irequency or other power source characteristics.

A) Goals (Continued)

The basic module shall be capable of compensating for fluctuations in primary line voltage Up to plus or minus 10% with no primary power limitation. Welding power shall remain constant within plus or minus 1%. Use of a closed loop system is desired in function.

Visual and/or audible signals shall be provided to indicate major power Supply module sub-system failure.

Each module or paralleled group of modules shall be capable of being controlled from the operator location. Each module shall be capable of automatic operation through the foot pedal control system.

Each basic module shall be rectangularly shaped to allow easy stacking during use and storage.

The design goal for the weight of a basic module is 40 lbs. A basic module includes no cables, brackets or attachments. Shipyard movement without the use of cranes and riggers is highly desirable.

Cooling shall be by convection and conduction.

The ambient temperature operating range of the basic *module* shall be from 0° to 120° F.

Duty cycle shall be 100%.

Maintainability:

The basic module shall not be open to the ambient environment. The module shall be free from vents and openings.

The basic module will not be water cooled.

The basic module shall not be cooled by ducted or forced air.

Power supply sub-system components shall be plug-in modules where practical.

Construction of the module shall be in accordance with good commercial practice. The module shall be capable of withstanding reasonably rough handling and transportation environments.

Design of the module shall provide easy access to the interior of the case to allow repair and replacement.

Materials:

The module shall contain supplies, products, or materials manufactured within the U. S. A.

B) Plan

The basic approach taken with this reacarch project is rather revolutionary. It is obvious that welding power supply operating from the 440 VAC 60 Hz line can never be made to meet the design goals discussed above, especially in regard to weight, duty cycle and cooling. Instead of using a conventional 60 llz transformer, a high frequency lightweight and efficient converter system is used. The basic concept is explained in Section II. For this reason, much of the research involved in this project is electronic in nature; therefore, much of the discussion in this report deals with the electronics of the proposed welding supply.

The method and plan of this research program is outlined *in* the following statement of work followed by the Celesco Engineering Staff.

1. Establish Design Goals and Prepare Conceptual Electrical Design

   a. Establish Design Goals

      Prepare documentation to guide the design and analyses of the portable power supply. This document will compile and combine desired portions of welding research project SP-1-2 (203.01) and Celesco proposal 0572-4-5.

   b. Conceptual Designs

      Prepare 2-4 feasible electrical circuits with block diagrams and schematics. Contact semi-conductor manufactrers and determine which state-of-the-art components can contribute to advanced designs. Review transformer core manufacturers' capabilities to provide advanced lightweight transformers.

   c. Select the Most Promising Circuit Designs

      (Formal Review)

2. Optimize Circuitry and Perform Assembly Tests

   a. Optimize Circuitry of Selected Designs

      Analyze each of the promising concepts for theoretical heat contribution of the power supply sub-assemblies. Utilize information on state-of-the-art semiconductors and circuitry to reduce heat contributions. Examine each subassembly for its weight contribution and optimize. This activity concerns primarily electrical circuitry and does not include designs of the case, receptacles, etc.

   b. Sub-Assembly Test

      Select promising circuits and sub-assemblies for breadboard type fabrication and test. Procure selected sub-assembly components, assemble and test. Determine heat contributions and ability to achieve various target weights.

(c) Review and Select Sub-Assembly Concepts Which Test Data Substantiates as Meeting Rcasonable Requirements

(Formal Review)

3. Fabricate Breadboards and Test

a. Prepare Candidate Circuitry Designs

Prepare circuit designs utilizing data from sub-assembly tests. Calculate thermal and mass properties of selected concepts. Determine control circuitry design and prepare to verify performance by incorporation in breadboard testing. Prepare breadboard test drawings of the power supply electrical system and prepare parts lists.

b. Assemble and Test Breadboard Electrical Systems

Procure and fabricate breadboard parts. Assemble the system to include sub-systems and test the power supply characteristics under a dummy load and while welding.

Provide system information to allow test prototype design.

Reduce power supply breadboard design features to a minimum in order to concentrate on the switching circuit with full line voltage.

Research adaptation of other industry power supplies and apply good features.

c. Review Results of Breadboard Testing

Review the results of the breadboard activity to arrive at the desired configuration for further development.

Establish the basic circuitry for the test prototype. Concentrate on the major design goals of power, control and reliability.

Continue the breadboard test and development concurrent with the evolution of a test prototype.

4. Prepare Prototype Design

a. Utilize Breadboard Test Data to Upgrade the Packaging Concept

Examine the on-going test results and identify proven approaches.

b. Prepare Power Supply Test Prototype Design (Preliminary)

Prepare preliminary power supply design. Provide a concept for case and heat sink configuration. Conduct preliminary thermal and mass characteristic analyses.

c. Determine Analytical Feasibility of Design Goals

(Formal Review)

5. Fabricate and Test a Test Prototype Power Supply

a. Prepare Final Prototype Design

Prepare final design of the prototype power supply. Provide multiple case designs if necessary. Conduct analyses to include design modifications. Prepare a parts list. Procure and fabricate parts. Assemble the test prototype system.

b. Prepare a Prototype Test Plan and Equipment List

c. Conduct Tests

Conduct tests of the prototype to determine its capability. Compare results with the design goals.

Modify the prototypes as required to approach design goal performance and maintain similarity to concurrent bread-board advancements.

Conduct a short period of controlled shipyard testing in San Diego. Record results.

Within budget and schedule constraints, the prototype unit will be tested and modified to approach design goals.

6. Documentation and Final Report

Prepare a final report which describes the chronological events of the program.

This report will include unsuccessful approaches, as well as successful.

Schematics, drawings, pictures and test results will be included.

II Basic Concept

A) Background

The primary source of weight in any standard power supply is the iron core transformer. These transformers are designed for operation at a 60 cycle line frequency. The efficiencies encountered at these frequencies require a large transformer that is capable of supplying the necessary power.

In the 1940's new power supplies appeared in use, primarily in military equipment, which were designed for operation on a 400 cycle line. As a result of the increased efficiencies thus obtained, these supplies were from 1/3 to 1/2 the size of their 60 cycle counterparts.

The supply to be developed under this contract was based upon a logical extension of this principle. The 60 cycle input power was to be rectified and converted to a DC level. This DC voltage was then to be switched at a much higher frequency. In the frequency range selected, from 5 to 10 KHz, a small toroidal transformer using a ferrite core can be used. The overall weight reduction to be realized is in excess of 10 to 1. An equivalent reduction in size is also obtainable.

B) Block Diagram

To achieve the results described, five basic building blocks are required. These blocks and their relationships are illustrated in Figure II-A

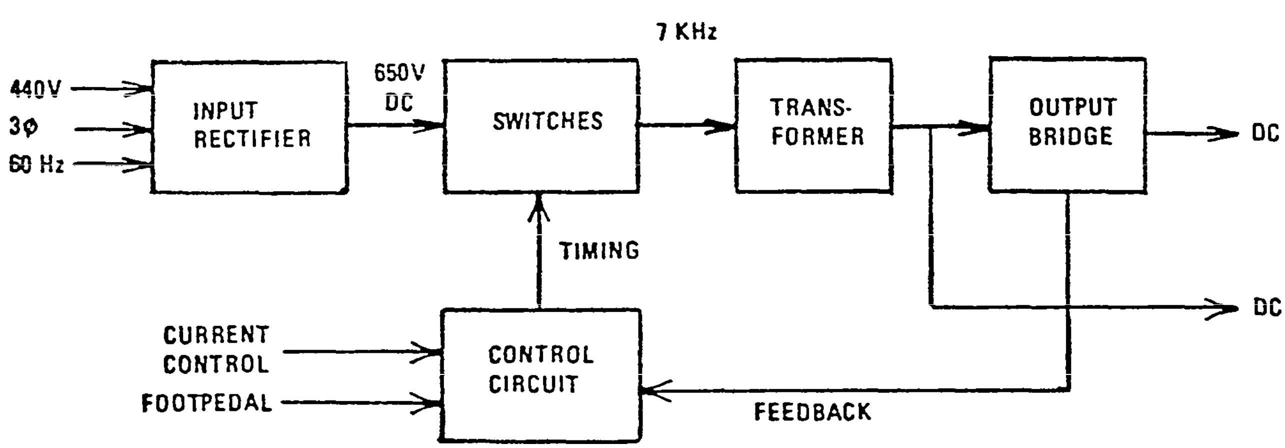


Figure II-A

B) <u>Basic Concept</u> (Continued)

As shown, the function of the input rectifier is to convert the 440V, 60 input to a DC level. This level is a function of the AC input voltage, in this case approximately 650 VDC.

The next circuit, the switch, converts the 650 VDC to a square wave alternating at 7 KHz at the same voltage level. This higher frequency AC voltage is then applied to the toroidal transformer where it is reduced to 20 to 40 volts as required, at a maximum of 300 amperes.

The control circuit performs three basic functions. First, it controls the action of the switches, turning them off and on in the proper relationship. Its second function is to convert input signals from the current control switch and from the foot pedal into switch control outputs. Its third function is to monitor the output current and hold it at the selected level despite variations in arc length.

The last block, the output rectifier, converts the transformer AC output to a pulsating DC for DC welding. Alternate output terminals apply the AC output directly from the transformer to the torch.

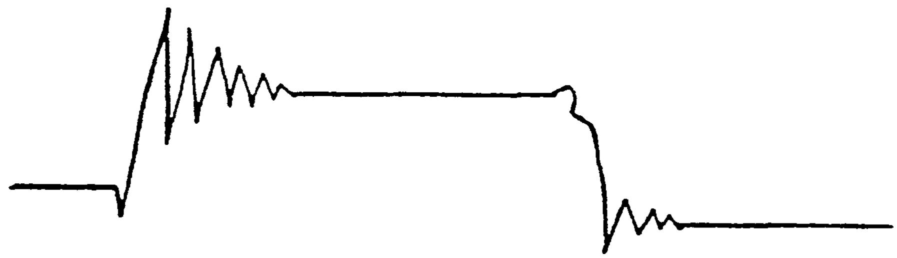
c) <u>Design Considerations - Electrical</u>

1. <u>Input Bridge Rectifier</u>

AC to DC conversion may be accomplished by using a standard 3 phase bridge circuit. Diode selection is based upon the voltage and current operating levels. Three of the six rectifiers can be replaced by SCR's to provide on/off capability similar to that of an output contactor on conventional welding machines.

2. <u>Switch Circuit</u>

This is one of the more critical areas of design. The switching components must operate at high voltage and current, as well as having the capability of very rapid switching. In addition, switching transients must be minimized to avoid destructive peaks. Switching components must be held in the on and off states with a minimum of time between. Since most of the power lost in a switch is lost during the switching interval, excessive switching times result in loss of efficiency and component overheating and possible damage. It can further be seen that switch efficiency improves as an inverse function of frequency since the switching interval becomes a proportionally small percentage of the overall period. Figure II-B illustrates a typical voltage waveform of a semiconductor device switching into a transformer load.



Voltage Waveform

Figure II-B

3. Output Transformer

The Output Transformer effects the energy conversion from a high voltage to a lower one at an increase in current capability. The use of ferrite core material is indicated to provide the necessary reduction in size and weight.

To provide a voltage output of 20 or 40 volts from the 650V input, transformer turns ratios of 32:1 and 16:1 were indicated; thus, a center-tapped primary is required. However, to compensate for output loading, ratios of approximately *one*-half the above amount will be used. Open circuit voltages of 40V and 80V will be generated, procducing the required levels under load. Transformer design calculations are presented in Section III-C-3.

4. Output Rectifier

The output rectifier section converts the transformer AC to to DC. A full wave rectifier will be used with a shunt in the common leg for output current detection. Since the output voltage levels are relatively low, the primary consideration is that of the current carrying capacity of the selected device and the speed of recovery.

5. Control Circuit

The control circuit must receive a reference signal from either a foot pedal or a selector on the machine itself. It must also receive a feedback (SMAW or TIG), *or* a voltage divider for a voltage process (MIG). The control circuit must compare the reference and the feedback signal and then drive the switches in a nccessary manner to achieve a zero difference or error signal.

5. <u>Control Circuit</u> (Continued)

The control circuit in conjunction with the power switches will provide variability. By utilizing some sort of pulse width control, the power to the transformer can be precisely controlled.

The control circuit must also provide some sort of current limiting so as to protect the semiconductors in the switches from overcurrent.

D) <u>Design Considerations - Mechanical</u>

1. <u>Size and Weight</u>

Keeping size and weight to a minimum are primary objectives of this program and are the major mechanical considerations in packaging the unit. Factors which determine size and weight are (1) total size and weight of required components, (2) size and weight of heat sinks required for adequate cooling, and (3) packaging concept.

Early in the program it was apparent that component quantity and size would not be the limiting factor as those selected for the early breadboard would readily fit into the required space. This left size and weight of required heat sinks along with packaging techniques as the major considerations.

2. <u>Heat</u>

A requirement of the system is that all cooling be natural convection in order to improve reliability and reduce maintenance problems associated with fans and blowing unfiltered air over heat sinks and components. At an efficiency of 90% the maximum output expected is 12,000 watts with 1333 watts to be dissipated. At 80% efficiency the dissipation would be 3000 watts. Final efficiency is expected to be bounded by these two limits. Using natural convection cooling, the major problem becomes available surface area on the exterior of the package. Through breadboard evaluations actual heat dissipation can be determined which can then be used in making the necessary heat sink calculations.

3. <u>Packaging</u>

Basic considerations for the packaging concept other than minimum size and weight are simplicity through a rugged functional design. The center of gravity and handles should be suitably located for ease of handling. The unit must be designed to withstand the abuse of shipyard environment. It must also be water tight in the final design and all exposed components be covered and insulated from the outside. The entire case and all heat sinks must be grounded to prevent hazard from electrical shock.

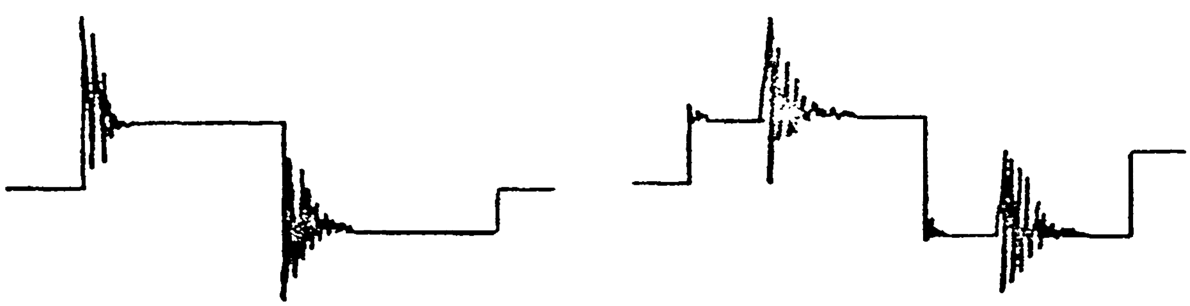
E) Problem Areas

1. Noise Transients

In the process of switching the large voltages and currents generated within the system, transients are generated which may be harmful to circuit components. These transients consist of momentary voltage or current surges which may greatly exceed the normal value. The use of reactive components in the circuit, such as the output transformer, may aggrevate the condition. In a system such as the supply in this case, where components are already being operated at their limits of current and voltage, this condition is particularly critical. The welding arc itself is a particular source of interference which must be guarded against. Another noise source is the power line itself, which supplies numerous welders of all types. Each of these sources may produce a surge which is capable of destroying the unit. Appropriate suppression techniques are required for each.

One common source of circuit noise is the transformer leakage inductance. In examining the relationships as represented by the equation $E = L\ di/dt$, it is readily seen that the voltage across the inductor is directly proportional to the size of the inductor and the rate of change of current with regard to time. Thus, switching 20 amperes in 1 microsecond results in a di/dt value of 20×10^6 amperes per second. The voltage generated by a 1 millihenry choke is then equal to 20,000 volts per second.

Some sources of switch transients and their effect on circuit waveforms are shown below in Figure II-C



a) Effect of energizing and de-energizing primary current.

b) Effect of sharply varying loads.

Figure II-C

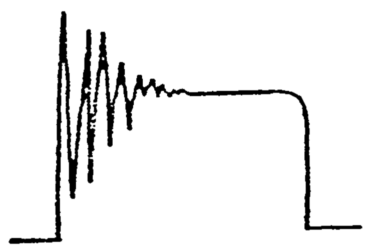
1. <u>Noise Transients</u> (Continued)

As can be seen from Figure II-C, abrupt changes, typical of normal circuit operation, may produce levels which are in themselves destructive and in addition which may produce a rate of change which exceeds component specifications.

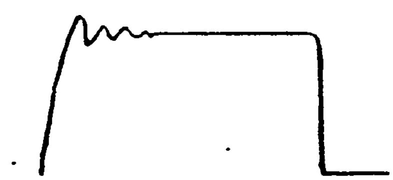
Additional problems are encountered in the responses of the semiconductors themselves. The stored charge may also produce undesired transients.

Suppression of these transients may involve a number of techniques. A final design will usually require a number of these.

The simplest method is that of the bypass capacitor. This capacitor, placed in parallel with a switching component, acts to absorb the undesired signal. (Figure II-D.)



a) Switching Transient



b) Effect of Bypass Capacitor

c) Overcompensation

Figure II-D

1. Noise Transients (Continued)

The use of bypass capacitors on the input bridge rectifier completely eliminated the catastrophic failures previously experienced in this area.

More elaborate filters may be composed of resistors, inductors or capacitors in combination. For example, a series R-C circuit was found to be required across the gate input of an SCR. Series inductors, with parallel RC circuits are also used to control commutation transients. It should be noted, however, that these circuits may, themselves, introduce undesired distortions which require further compensation.

Another method of transient control is by means of voltage limiters. The two basic components of this type are zener diodes and varistors. Both of these components display the characteristic of having no effect upon a circuit until a specified voltage level is exceeded. At this time, the units conduct, absorbing the undesired peak. In the lightweight power supply, varistors are placed across each transistor switch for over-voltage protection. Zener diodes are used in the transistor base circuits for the same purpose.

2. Component Ratings and Availability

A basic requirement of the switching circuit is the ability to switch up to 30 amperes at 650 volts. Although a number of transistors were available at the beginning of the program which were capable of switching the desired current, none could do so at the required voltage. Thus, it became necessary to parallel a number of transistors of lower rating to meet the current requirement. This problem is aggravated by the fact that the current handling capabilities of transistors must be heavily derated as the operating voltage is increased. The specific components selected, and the trade-offs involved, will be discussed.

III Areas of Investigation

A) Semiconductor Switch Background

As was noted in the previous section on problem areas, the most significant obstacle posed by this project is the switch circuit. The conversion from 650 VDC to 7000 Hz AC must be accomplished by semiconductor switches in order to maintain efficient lightweight conversion.

There are two types of semiconductor devices that can be used in a lightweight power conversion system; the Silicon Controlled Rectifier (SCR) or the Silicon Transistor. In addition to these two devices, there are numerous configurations in which the two can be operated. The optimum device and configuration must be chosen to minimize weight and maximize efficiency.

In addition to the problems mentioned in the previous section on noise and component availability, there is also problems involved in energy dissipation. Two types of dissipation are:

1. Saturation dissipation - power lost during the on state of a device. SCR's have an inherently higher saturation voltage than do transistors, and thus have higher saturation losses.

2. Switching dissipation - power lost in the transition from the on to off state. Transistors have an inherent problem of saturation delay time; this causes higher switching losses.

B) Transistor Configurations

The basic inverter configuration used for transistor switching is the bridge configuration, since transistor voltage ratings prevent a center - tapped configuration. Therefore, only the switch design is discussed in the following discussion on transistors. Keep in mind, four sets of switches are used even though only one set is discussed.

Serial Transistors

Initial investigations into inverter operation used Delco transistor type DTS 721 in parallel banks. The maximum collector rating of 800 volts appeared to allow sufficient margin for satisfactory operation at the required 650 volt level. Since the transistor is capable of switching only approximately 3 amperes at this level, it was necessary to parallel ten units to provide the desired current capability. However, in attempting to operate at these levels, extensive component damage was realized from reiatively minor overvoltage conditions. Due to this, it was decided, early in March, to investigate the possibility of replacing each single transistor with a series pair. This combination reduces the voltage appearing across any one transistor to one-half the former value. Other benefits were realized.

Serial Transistors (Continued)

The greatly reduced voltage allowed the transistors to be operated at higher current ratings. Further, more transistors were then available for operation at this voltage having even higher current ratings. The transistor selected *for this* purpose was the RCA 2N6251. This permitted the number of parallel units to be reduced from 10 to 3. The switch configuration shifted then from 10 transistors in parallel, to three parallel pairs of transistors in series, as shown in Figure III-A.

Early results were promising. However, continuing tests revealed the difficulty in turning off both transistors in a pair exactly at the same time. Even small differences caused the entire voltage to appear across a single transistor, resulting in its destruction. (See Waveform Figure III-B.)

Some improvement was made in this condition when the driver transformers were rewound, exercising particular care to insure symmetry and uniformity; however, small differences persisted. (See Fig. III-C.)

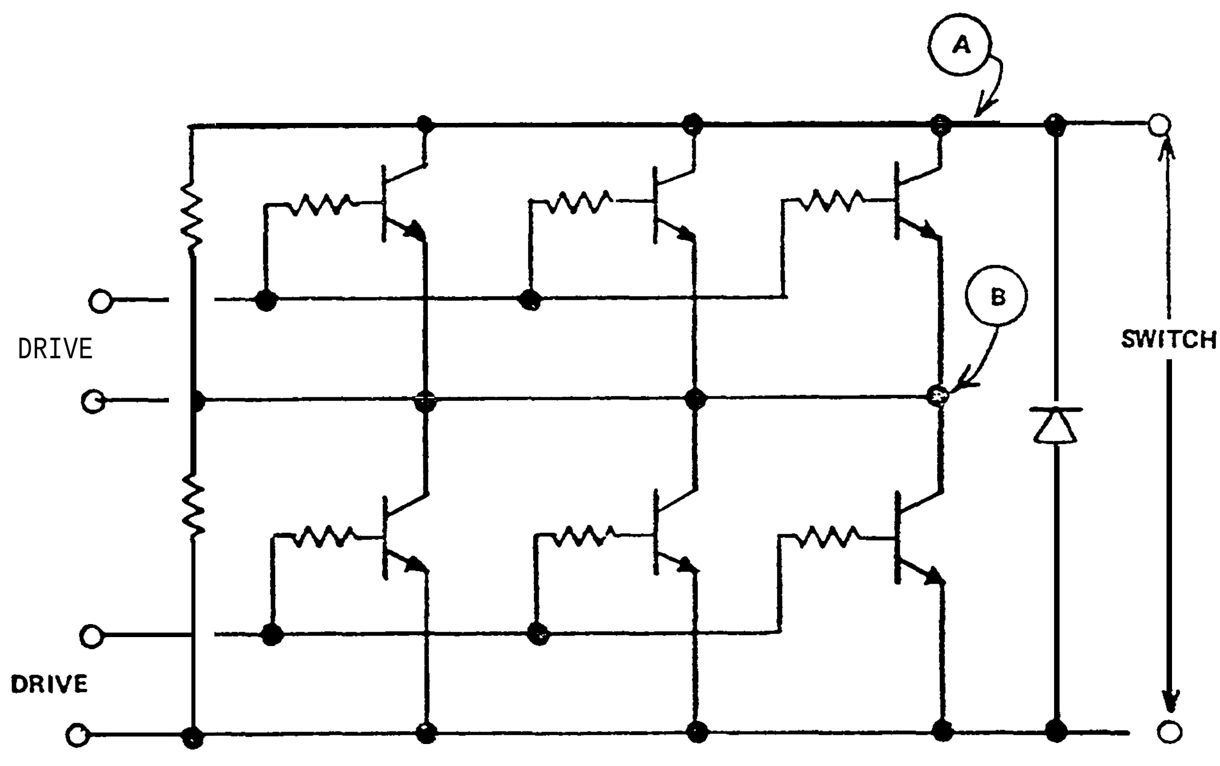
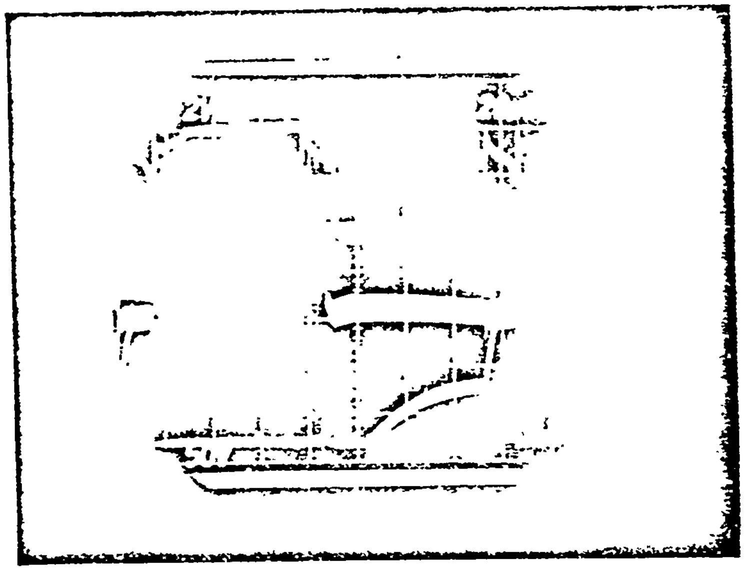


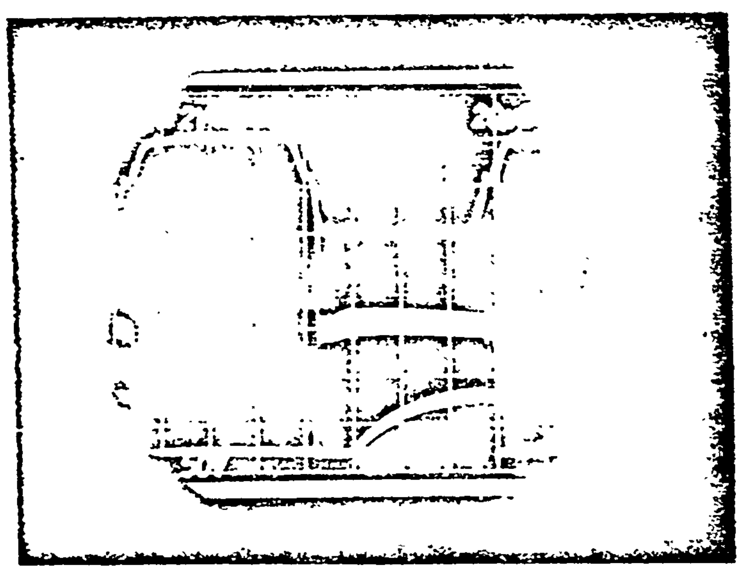
Figure III-A



Scale:
Vert:
Upper - 2V/CM
Lower - 50V/CM
Horiz: 20μ sec/CM

Figure III-B

Upper trace - Base drive
Lower trace - voltage waveforms at B (Figure III-A) (square) and A (Figure III-A). Note at the beginning of the rise, full voltage is across upper transistors, since the voltage across the lower transistor remains zero and only slowly rises to half value.



Scale:
Vert:
Upper - 2V/CM
Lower - 50V/CM
Horiz: 20μsec/CM

Upper trace - base waveform - note improvement over Figure III-B
Lower trace - collector waveforms - note improvement, but still, full voltage appears across the upper transistors at the rise period.

Figure III-C

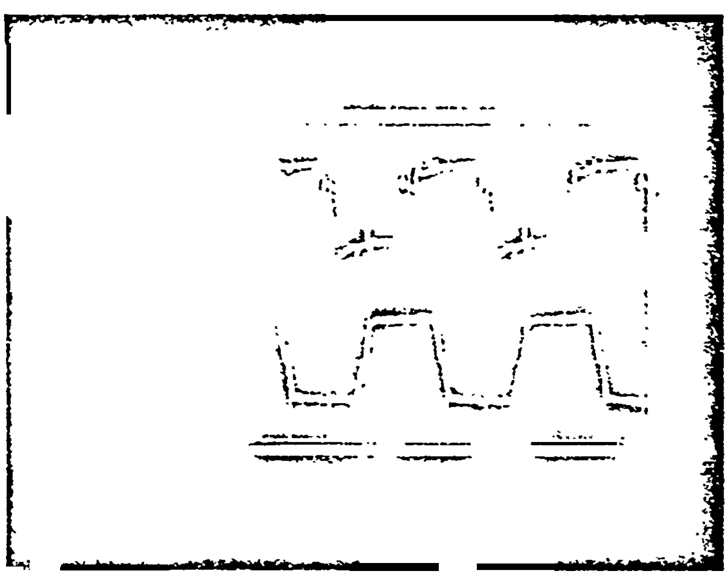
Serial Transistors (Continued)

Capacitor and inductor equalizing networks were installed across the series pairs with further improvement. Although DC biasing of the higher level transistor was considered, to insure automatic cutoff when the lower one was switched off, this approach was discarded due to the excessive heat and power wasted in the bias circuit. Analysis of test data indicated that this system could, with care, be refined to produce a supply of the required capacity.

Parallel Configuration

Since a number of new transistors having higher voltage and current ratings had appeared in recent months, a new literature search was accomplished in April, 1973. As a result, several new transistors were obtained and evaluated One of these, the General Electric D56W2, appeared promising, being rated at 1400 volts at 5 amperes under the required operating conditions. However, this transistor was found to have such a low gain as to make it useless for this application.

Another transistor, the Delco DTS 804, was obtained shich combined the higher voltage and current ratings with an acceptable gain figure. The maximum collector voltage rating of 1400 volts 800 volts SUS. offered a considerably improved safety margin. Since this eliminated the necessity for the more complex series pair, the switch configuration was redesigned using the DTS 804 in parallel banks of ten. As shown in Figure II-J this is the configuration selected as being the most likely to result in an operating supply. A redesign of the control transformers was required to match the new transistor characteristics. Figure II-D illustrates the improvement in waveform realized by this procedure. The system operated clearly at lower power levels; however, as the input voltage was increased, voltage transients appeared in the drive signal. Additional filtering of drive and switching circuits eliminated this condition as shown in photos Figure III-E and Figure III-F.



Scale:
Horiz: 20μ sec/CM
Vert: 2V/CM

Figure III-D

Upper trace - old drive transformer - note the unsymmetrical shape about the centerline of the waveform.
Lower trace - improved drive transformer.

Scale:
Horiz: 20 μsec/CM
Vert: 2V/CM

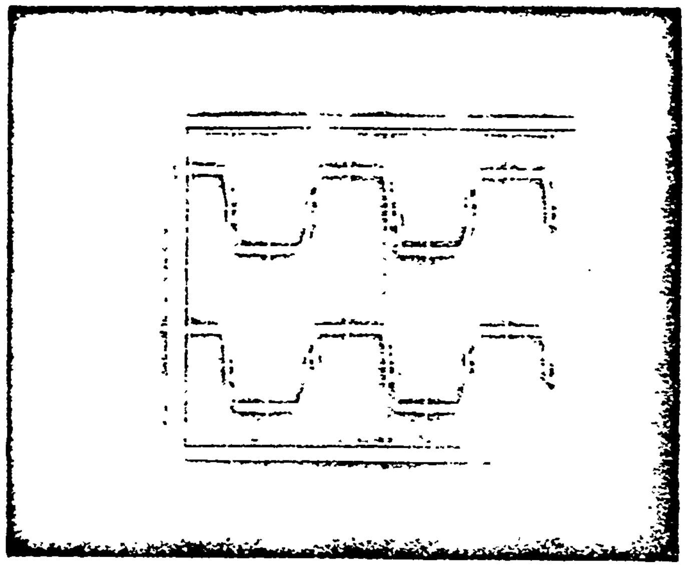


Figure III-E

Two waveforms represent base drive on opposite switches.

Scale:
Horiz: 20 μsec/CM
Vert: 2V/CM

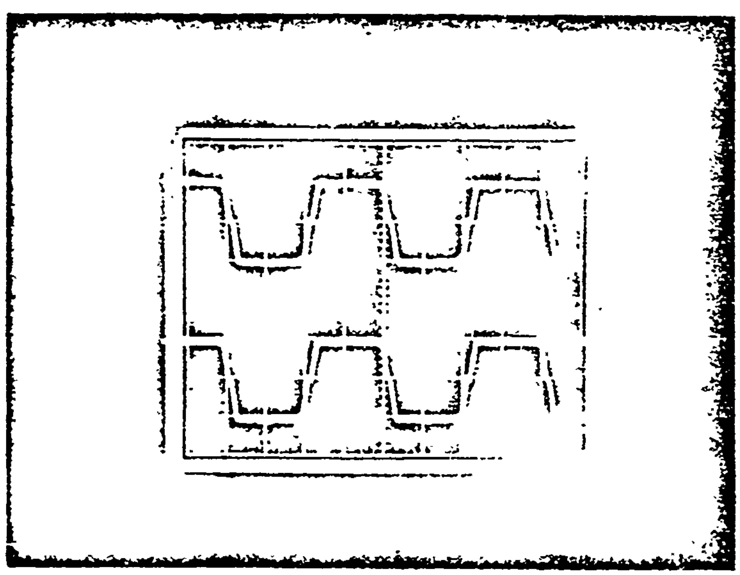
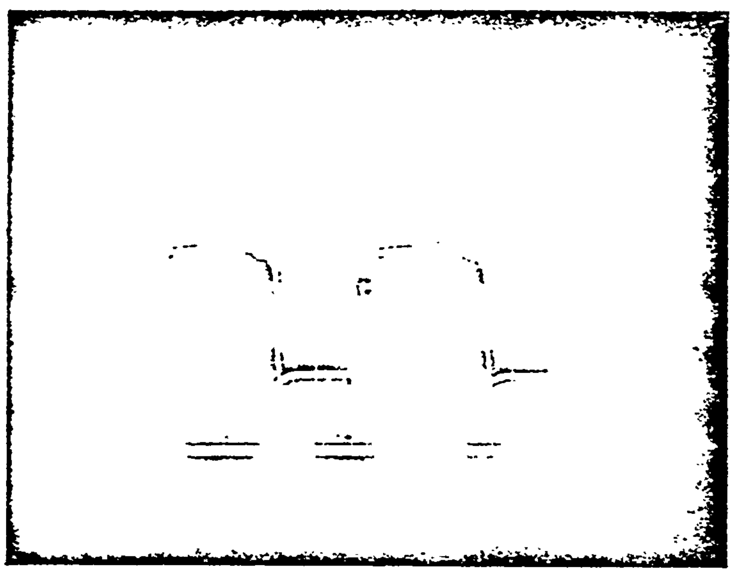


Figure III-F

Serial Transistors (Continued)

Early calculations indicated that base drive requirements would be critical, both as to supplying sufficient drive to hold the switches in the on state at high power levels and to quickly eliminate the stored charge for rapid turn-off. In the month of April, as the power supply began to be operated at higher power levels for longer periods, additional base drive problems were revealed. Tests were made during this time to evaluate the effects of base drive levels and current impedance on switch performance at higher power levels. As anticipated, increasing the base drive without additional compensation resulted in a slower collector turn-off. A typical collector waveform at NO levels of base drive is shown in Figure III-G and Figure III-H.



Scale:
Vert: 50V/CM
Horiz: 20μsec/CM

Figure III-G

Note turn-off delay.

Scale:
Vert: 50V/CM
Horiz: 50 sec/CM

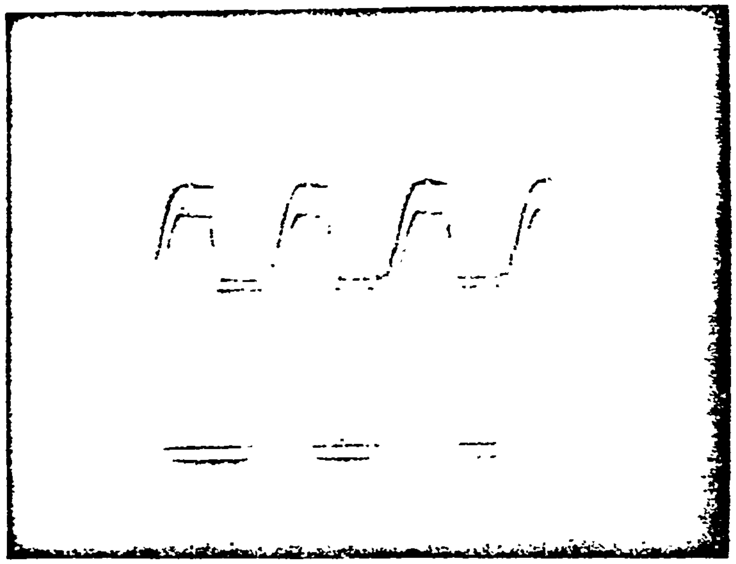


Figure III-H

Modification of the base circuit of the switch drivers corrected this condition as shown in Figure III-I. This photograph presents the composite waveform of two transistor switches at low ($\measuredangle$100 a) current levels. This configuration was then run extensively at low power levels with satisfactory results.

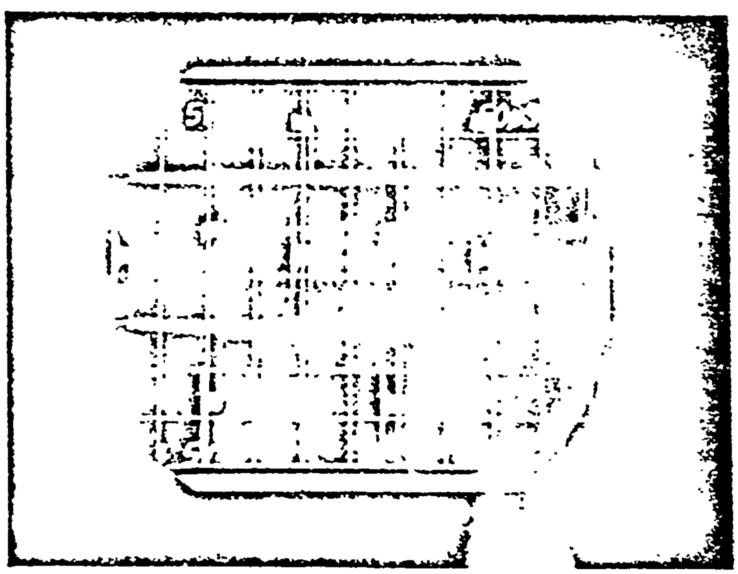


Figure III-I

Serial Transistors (Continued)

A new problem was encountered at high power operation. As shown in Figure III-K, a phase stift was encountered between the primary and secondary of the power transformer which did not appear at the lower levels. As is shown in the photo by vertical line A, the leading edges of the secondary low power pulses are coincident in time with those of the primary at time B, a delay is seen between the leading edges of the primary and secondary. Analysis of this condition indicated this condition to be the result of unbalance existing in the power transformer. A new transformer was wound which corrected this shift. Photos of low and high power output signals showing the resulting symmetry and transient free waveforms are presented in Figures III-L and III-M.

Scale:
Vert: 50 V/CM
Horiz: 50μsec/CM

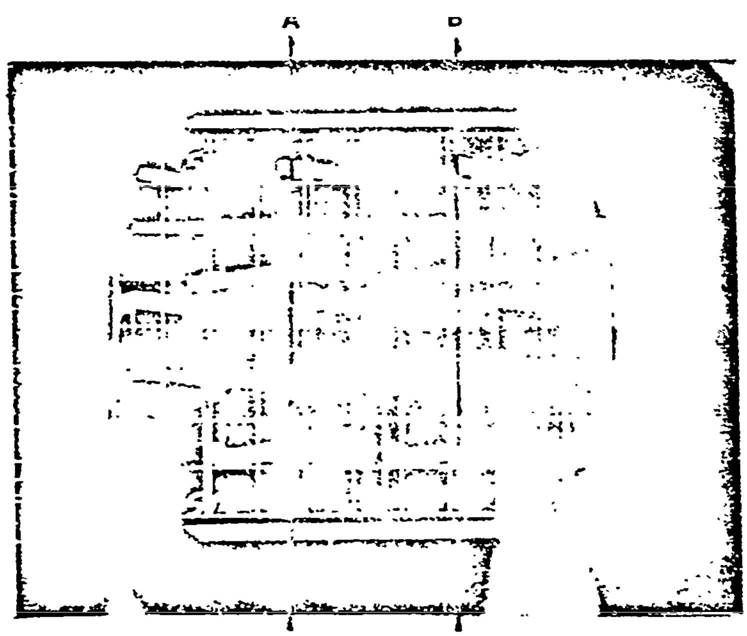


Figure III-K

Scale:
Vert: 50V/CM
Horiz: 20 μsec/CM

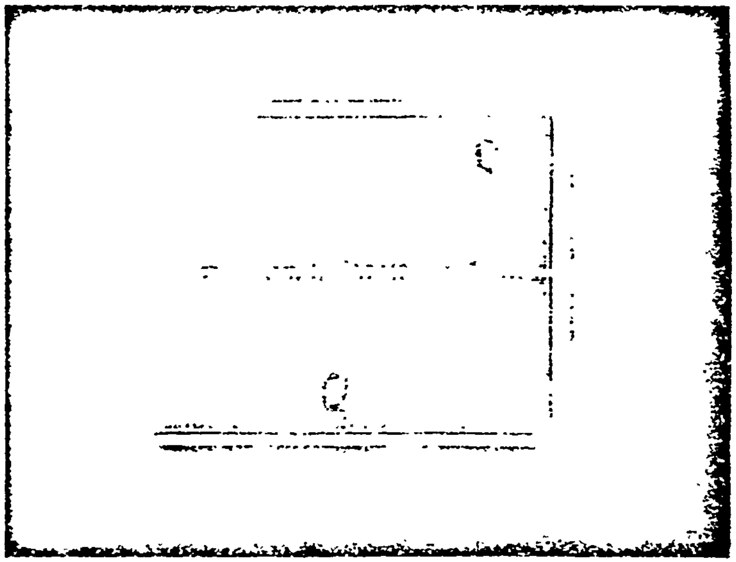


Figure III-L

Low power output - note narrow pulse width.

Scale:
Vert: 50V/CM
Horiz: 20 μsec/CM

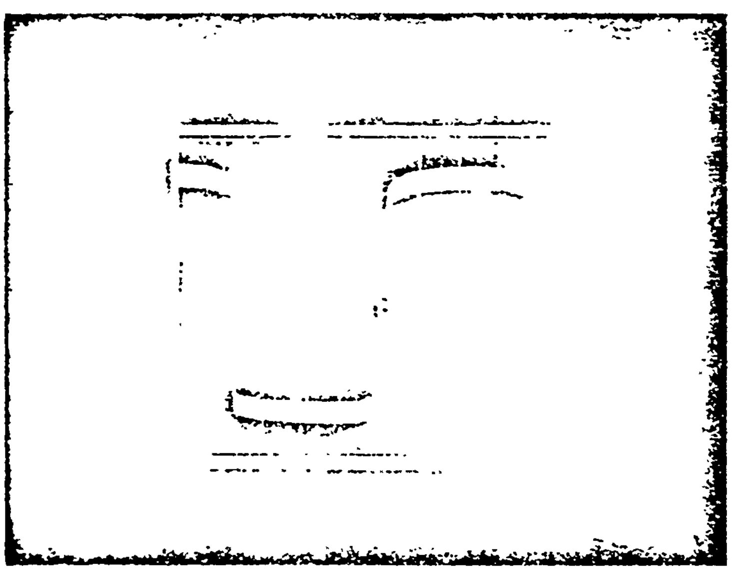


Figure III-M

High power output - note wide pulse width, almost square.

c) Silicon Controlled Rectifiers

Unlike the transistor, the SCR's only function is that of a switch. The SCR is turned on by a current pulse to the gate lead; the device then conducts anode to cathode. The main disadvantage of the SCR is that it is not as simply turned off; the SCR is turned off only by some other component that either bypasses or interrupts the current in the SCR. Most commercial applications involve the AC line, and the alternating current is used to turn off or commutate the SCR.

The heart of the investigation of SCR's consisted of solving the commutation problem since SCR's are available with high voltage and current rating. In all cases, however, inverter type SCR's were chose due to their higher frequency characteristics. The SCR used in the subsequent investigations was the GE Cl59PB.

The first SCR circuit in'estigated was the coventional capacitor commutated circuit shown in Figure III-N. This circuit incorporates line commutated SCR's as an input switch. Two inverter SCR's are used in a certertapped inverter configuration.

One of the basic problems of this circuit is that in order to commutate one SCR off, the other must come on. This does not allow for variability of welding current or voltage. The line commutated SCR's were *then* attempted to be used as phase control SCR's, thus varying the DC voltage to the inverter. Through calculations, it was observed that the filtering required between the line SCR's and the inverter would be quit heavy. This is due to the 60 Hz frequency of the AC line.

This SCR circuit was operated successfully under load; the waveforms are shown in Figure 111-0. However, under no-load conditions there was no current to cause the commutating capacitor to recover. The result was that both SCR's would fail to commutate, drawing excessive current, resulting in their destruction. The only alternative was to use a dummy load to permit the capacitor to recover. This idea was abandoned due to the excessive heat that would be dissipated in the dummy load.

Since open circuits exist continually in a welding operation, this circuit was considered unusable, not only because of its failure to commutate properly, but because of its inability to result in a variable output.

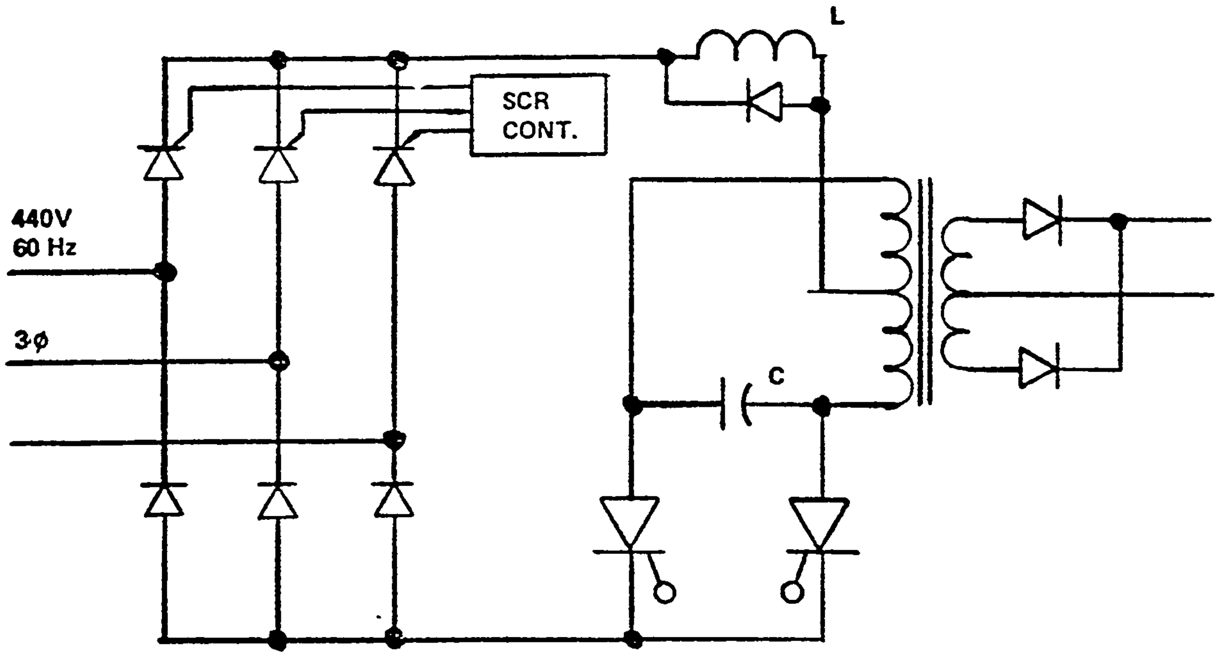


Figure III-N

Upper - 50V/CM
Lower - 20 ma/CM
Horiz: 50 μsec/CM

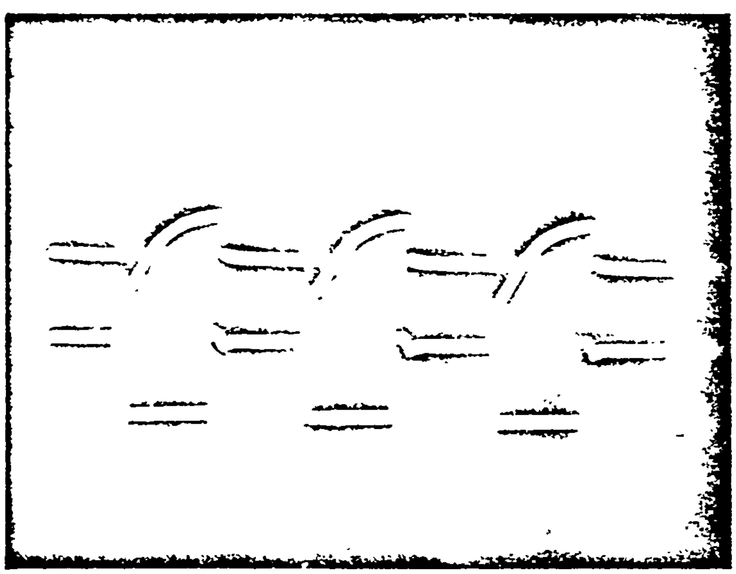


Figure III-0

Upper trace - SCR voltage note the negative spike which commutates the SCR and the exporential capacitor recovery.
Lower trace - gate current.

c) <u>Silicon Controlled Rcctifiers</u>

An SCR circuit which alternately charged and discharged a commutating capacitor through the power transformer primary was attempted. The basic schematic is shown in Figure III-P.

This curcuit was operated with some success, but the output power and efficiency can only be controlled by the frequency of operation. The charge and discharge of the capacitor again depends on the load of the transformer. The frequency of the two switches must be below the charge time in order to effect dependable commutation. The variable frequency is felt to be undesirable for welding applications. Efficiency is excellent, however, assuming adequate control circuitry is developed to provide the variable frequency needed to assure commutation.

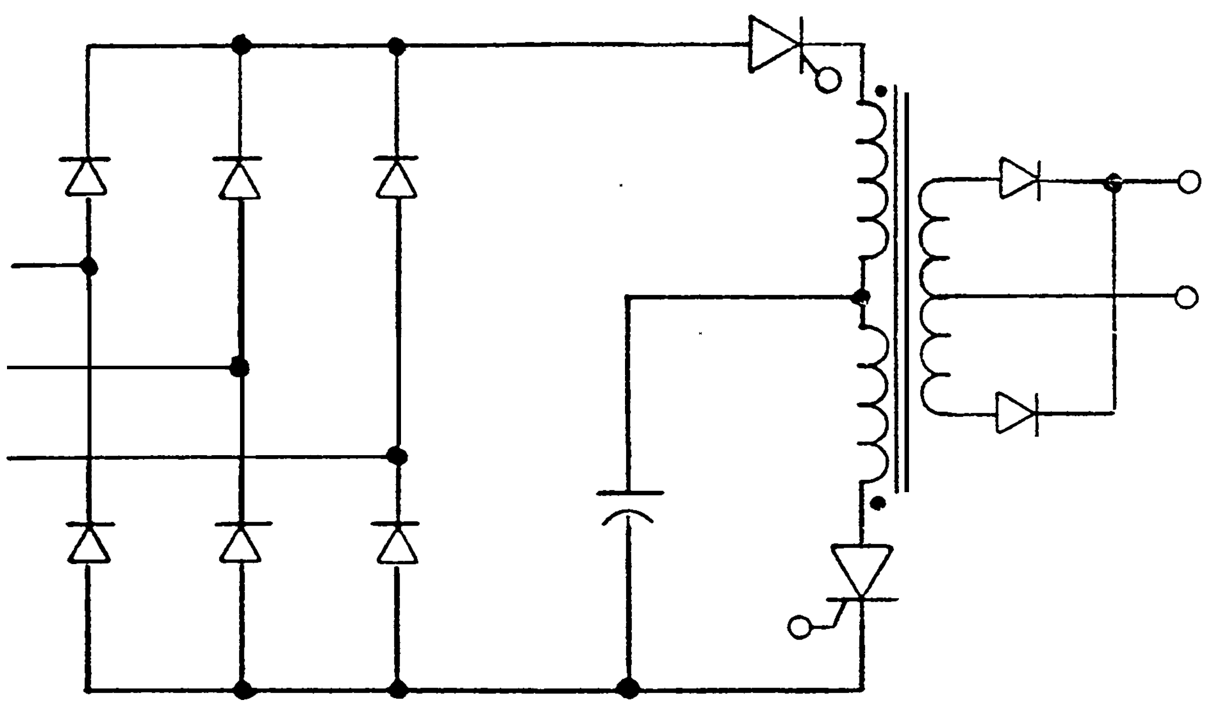


Figure III-P

c) Silicon Controlled Rectifiers

It was decided that reactive components could not be used to effectively to commutate the SCR's since the supply would be subjected to sharply varying loads. An SCR transistor hybrid circuit was developed in order to commutate the SCR's without using reactive components. The basic circuit is shown in Figure III-Q.

This circuit showed good promise; however, because of the transistor switch, this circuit had some of the problems and disadvantages associated with the transistor circuit. The problems were somewhat intensified by the voltage and current transients caused by the SCR switching. The diode DI was used to limit excessive voltage transients appearing across the transistor switch. This diode, however, impaired commutation. When the transistor switch interrupted the current I_1, the energy stored in the field in the transformer caused current I_2 to flow through the diode D and SCR to be commutated, thus delaying the commutation of the SCR.

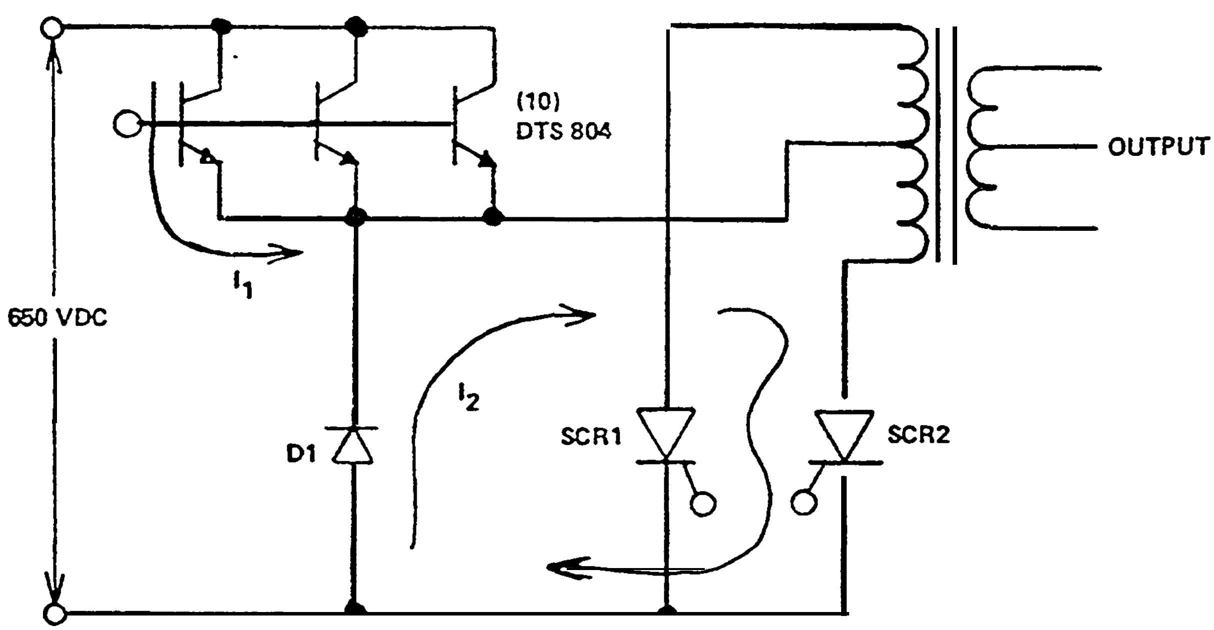


Figure III-Q

c) Silicon Controlled Rectifiers

Various other commutation schemes were investigated at lower power levels. A summary of the results are presented in the following discussion. One manufacturer has broken down the various methods into six basic categories. For ease of discussion, this classification will be used in this report

Class A

Class A commutation depends upon resonating the load to bring about S CR turnoff. An LC resonant circuit is used for this purpose which carries the full load current. For this reason, the resonant self-commutation is most suited to frequencies above 1 KHz. The output current is nearly sinusoidal with a corresponding low initial di/dt.

Characteristics of this circuit were found to be good output waveform with excellent load regulation. It operates well into an open circuit, as well as a variety of reactive loads. Some problems were encountered with operating this system in a variable pulse width mode, due to the resonant circuit requirements.

Class E

This method is similar to the above in that it uses an LC circuit to achieve commutation. Regulation is accomplished by time ratio control. This circuit is of limited use, one example being the Morgan chopper. Some problems encountered were circuit instability and transients resulting from the reactive components. Regulation was difficult to achieve, and excessive sensitivity to load fluctuations was noted. Although these limitations can be overcome through the use of appropriate techniques, the approach was found to be generally unsatisfactory for the desired applications.

Class C

This is the simplest of the commutation systems, using a capacitor or L-C circuit switched by the load carrying SCR. The best known example is the McMurry-Bedford Inverter. Optimum values of L and C are derived from the following equations:

$$C = \frac{T_c \, I\,com}{1.7E_b}$$

$$L = \frac{t_c \, E_b}{.425\, I\,com}$$

In addition, an Ott filier may be used to elimainate harmoliics. Good regulation may be achieved with this circuit; however, due to the L-C commutation components, this circuit is best used at frequencies

Class C (Continued)

below 1 KHz. In practice, difficulties were encountered in maintaining commutation and in achieving the desired efficiency and regulation.

Class D

In this method, an auxiliary SCR is used to switch an L or LC circuit. High efficiencies may be achieved with this circuit, as the commutation energy may be transferred to the load. Duty cycles of up to 80% may be realized. Puise width modulation is readily achieved, and the circuit works well at high frequencies. One example of this approach is the Jones chopper. The test circuit operation agreed closely with the literature and with predicted figures. This circuit was considered to be one of the more promising for use on the welding power supply.

Class E

Class E commutation is achieved through the use of an external turn-off pulse. Although this circuit is not widely used, due in part to the requirement for additional components, it is capable of high efficiencies and is readily adaptable to the pulse width modulation system selected for use in this application. Some progress was achieved in operating a pulse commutated system.

Class F

This is the most basic of SCR commutation systems, in which the normal operation of an AC line as the line voltage in changing polarity automatically results in turning off the SCR. Since it depends on the AC input for commutation, its use is limited. For this application (after rectification of the 440 VAC 60 Hz 3Ø line), there is no line; therefore this method is not applicable.

D) Comparison: SCR's Vs Transistors

Since reasonable success was achieved with transistor switches, it was decided that the transistor bridge circuit offered the best alternative for achieving a demonstrable unit at the termination of this project. However, it should be noted at this point that if SCR control problems could be solved, a more economical, simple design would result. To provide a timely demonstration of the feasibility of the high frequency approach, the full effort was directed toward an all transistor unit.

Some of the advantages and disadvantages of the SCR's and transistors *are* tabulated in Figure III-R. There are only two disadvantages with SCR's: lack of l-amiability and lack of adequate switch control. These two disadvantages diminished considerably the possibility of using SCR's as the switching elements.

D) Comparison: SCR's Vs. Transistors

As can be seen in Figure III-R, the transistor has more disadvantages than the SCR, four total:

1. Adequate ratings
2. Complexity of design
3. Difficult switching characteristics
4. Lower dissipation characteristics

All of these disadvantages are tolerable (although highly undesirable). The transistor, on the other hand, offers nearly complete variability and switch control, two necessary characteristics the SCR did not inherently have. For these reasons, the transistor was selected as the device most capable of achieving a demonstrable unit.

	SCR's	TRANSISTORS
1. Adequate component ratings and availability	Adequate ratings of SCR's are available to provide	Highest available rating (DTS 804 Transistor) provide for only 18% safety factor
2. Simplicity of circuit design	Two SCR's are needed for a centertapped inverter	Forty DTS 804's are needed for a bridge type inverter
3. Variability of power level	The SCR centertapped inverter does not provide for variability	The transistor bridge inverte provides pulse width variabili The current level (output) cou be varied from 20 to 250 amps
4. Efficiencies and switching loss	If commutated properly, SCR inverters can attain very high efficiencies	Due to the charge storage phenomena of transistors, it is difficult to attain efficiencies in the 90% due to the switching loss incurred
5. Dissipation capabilities	The SCR is capable of operating in some cases to 160°C	The transistor must be heavil derated at 75° and is useless at 100°C (DTS 804)
6. Switch control	The SCR is difficult to commutate especially under sharply varying loads	The transistor can be turned on and off by the control signal at the base

E) Control Circuitrv Design

In order to drive the switching circuits in the correct phase and duration, control signals must be generated and conditioned by buffer drivers. It is also necessary to provide a means of selecting the maximum current and to provide for control by a foot pedal. Included in this section is the feedback curcuit which senses the supply output current and maintains it at a constant level, regardless of variations in arc length.

Operational Amplifier Control System

The first approach to switch control utilized operational amplifiers connected in several configurations to generate the necessary control functions. This circuit operated as required, however, some instability was noted in the frequency control section. A complete circuit schematic is presented in Figures III-S to III-V.

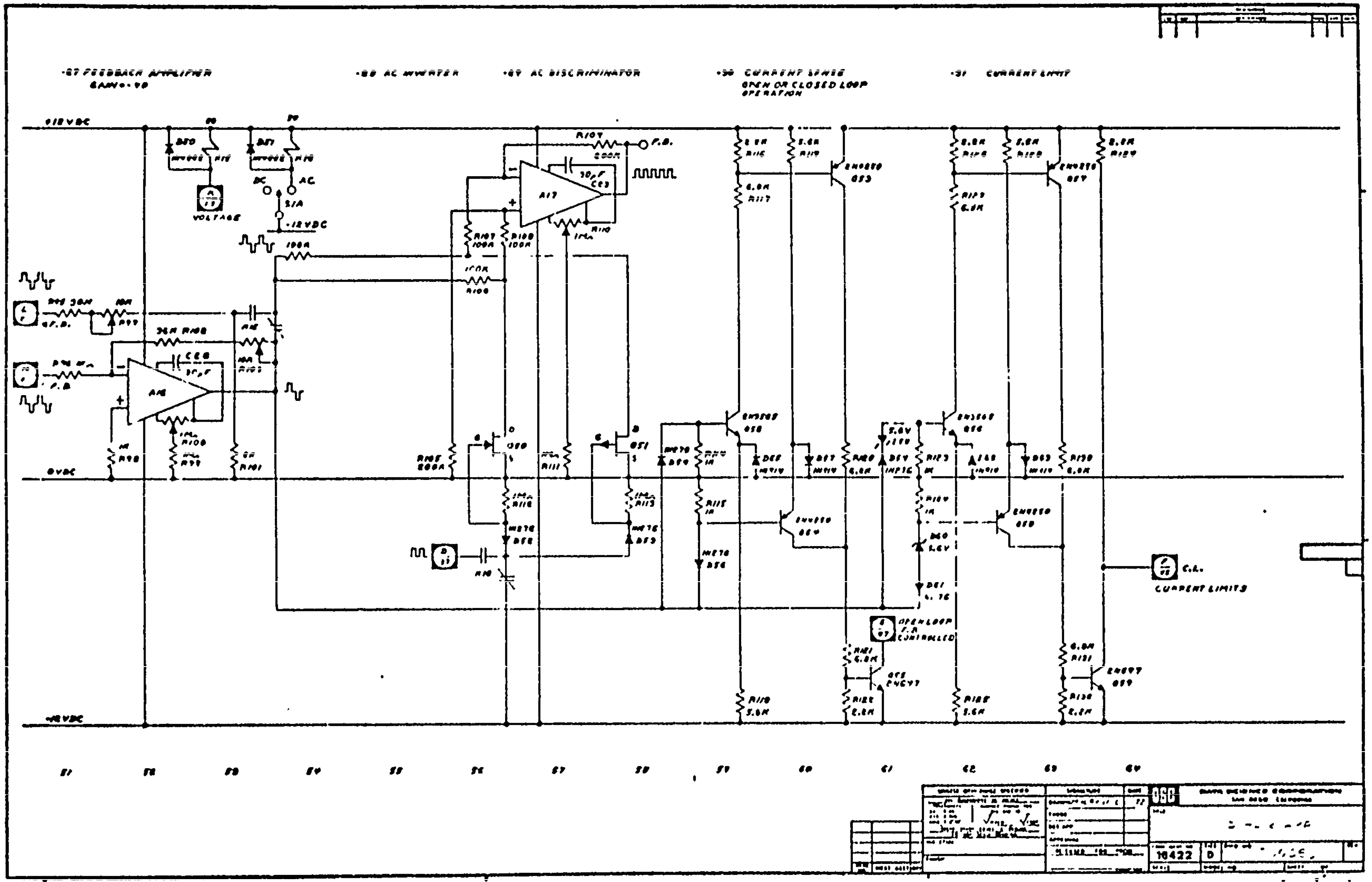
Digital Logic

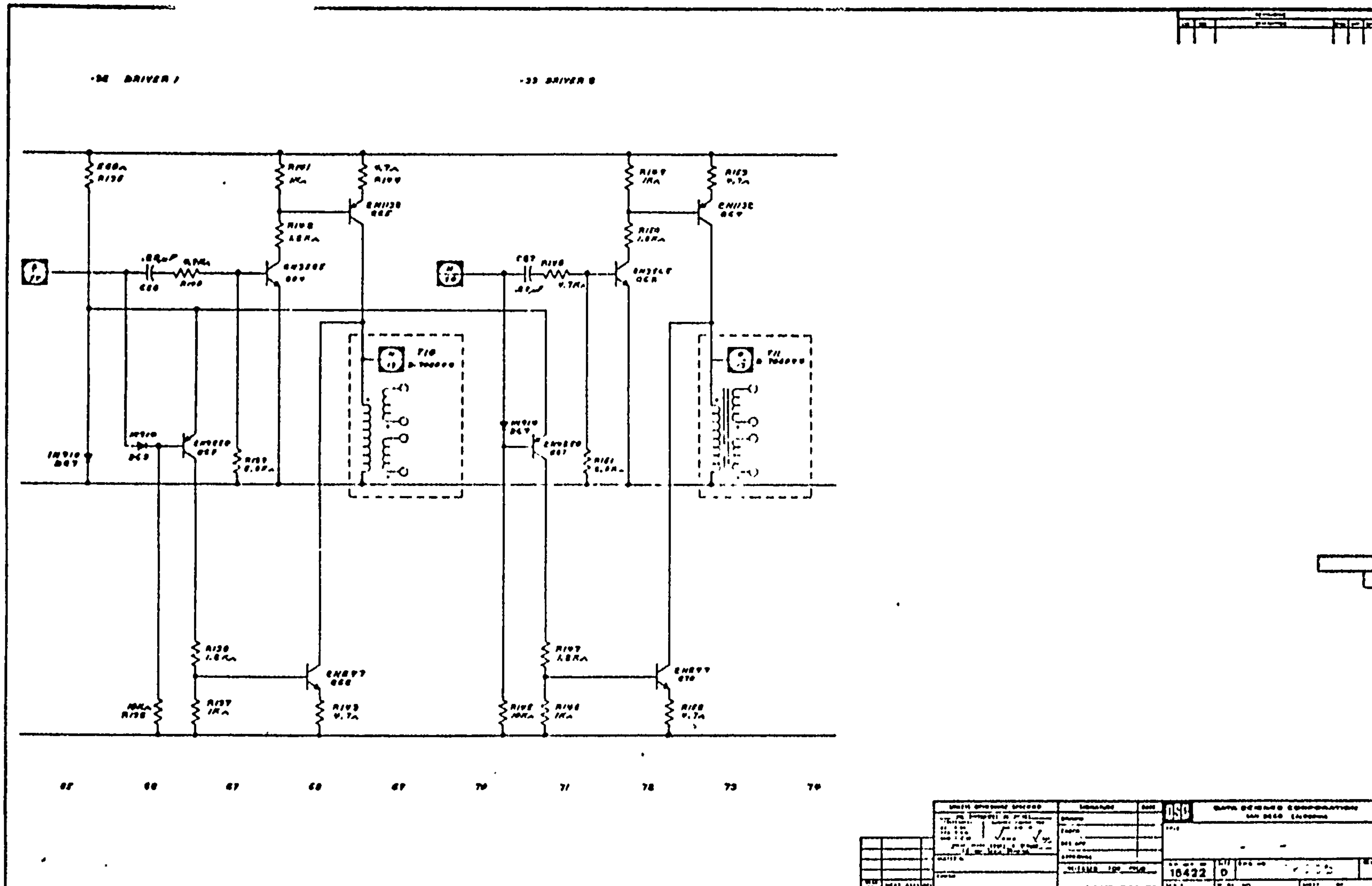
Due to the stability problem and to the inherent complexity of the operational amplifier system, a digital logic circuit was developed which performed the same function. This approach, due to its simplicity and reliability, was selected for use in the final power supply configuration. A schematic diagram of the final circuit is presented in Figure IV-E and IV-F of the next section.

Phasing Considerations

The first switching circuits utilized four transistor switch sections arranged in a bridge configuration. This system requires that alternate circuits be switched on simultaneously to provide a current reveral in the transformer primary. The period during which both switch sections are turned on, and during which the transformer primary is drawing current in one direction, is a variable which controls the amount of current available at the output. Although this circuit works well in terms of current control, some problems were encountered during the quiescent, or non-switching period. Due to the fact that the primary of the power transformer was left floating during this interval, excessive transients and waveform distortion were produced. To correct this condition, a modified switch cycle was implemented which maintained the transformer either in an energized or a grounded condition at all times. Original and modified sequences are presented in Figure HI-W a and b.

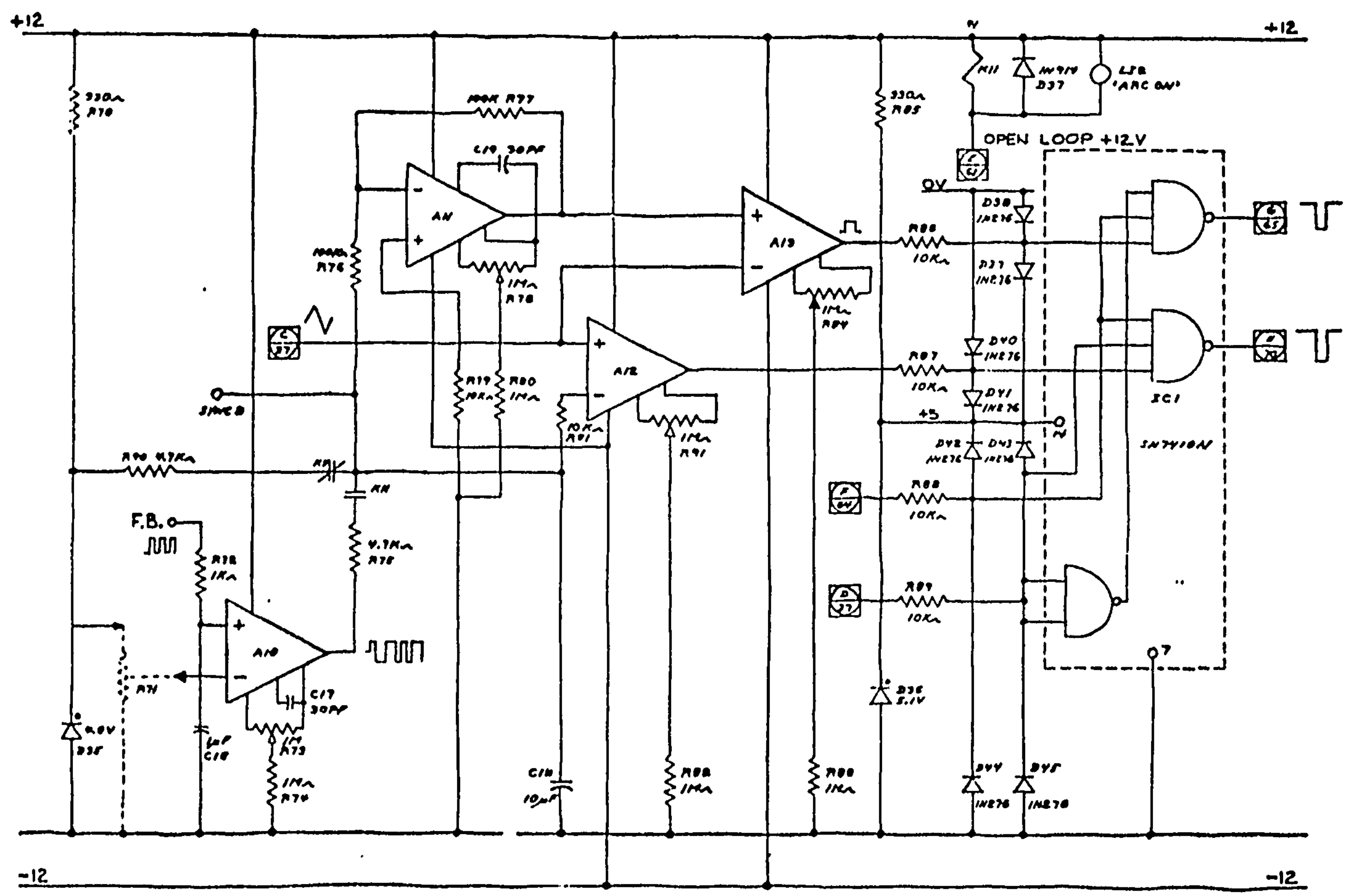
Further problems were encountered in achieving correct switch timing due to transistor turn-off time. Referring to Figure III-W-a, it can be seen that if switches 1 and 2 or 3 and 4 are turned on at the same time short circuit currents will flow, resulting in extensive component dam
As a first approach to the problem, careful attention was paid to the base drive circuits, insuring that the stored charge on the bases was quickly swept out. The effect of excessive turn-off time is shown in Figure III-X.


FEEDBACK AMPLIFIER
AC DISCRIMINATOR
CURRENT SENSE
OPEN OR CLOSED LOOP OPERATION
CURRENT LIMIT
+12VDC
0VDC
-12VDC
VOLTAGE
CURRENT LIMITS
10422



PULSE WIDTH MODULATION

8



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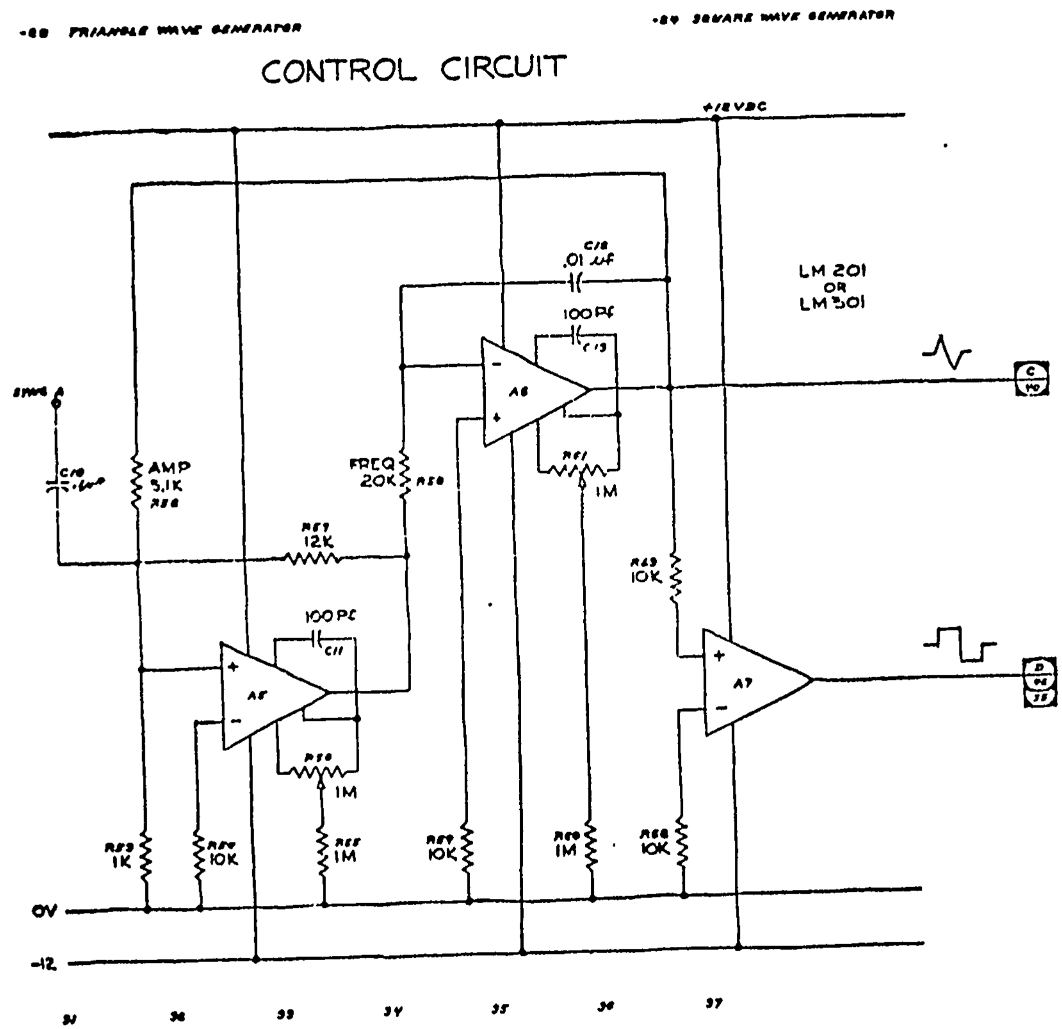
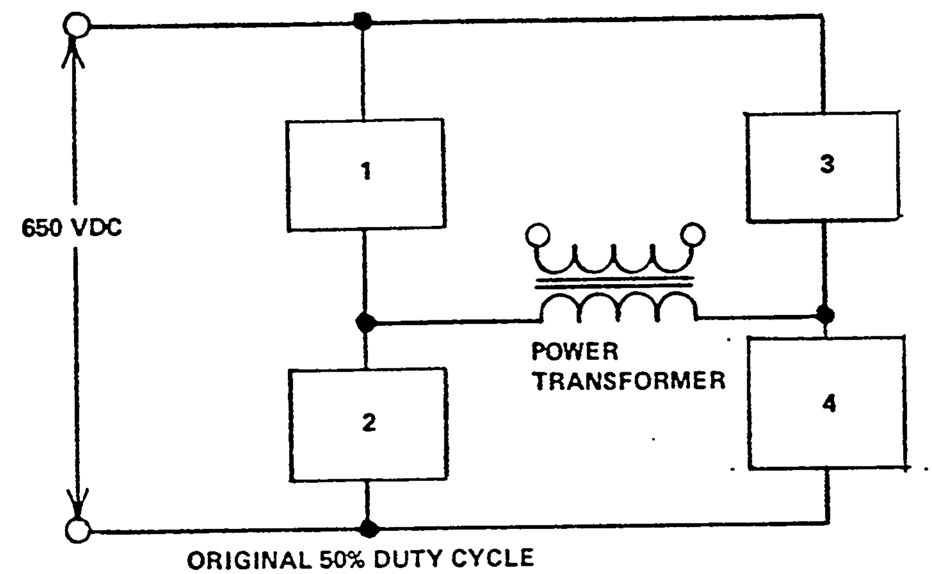
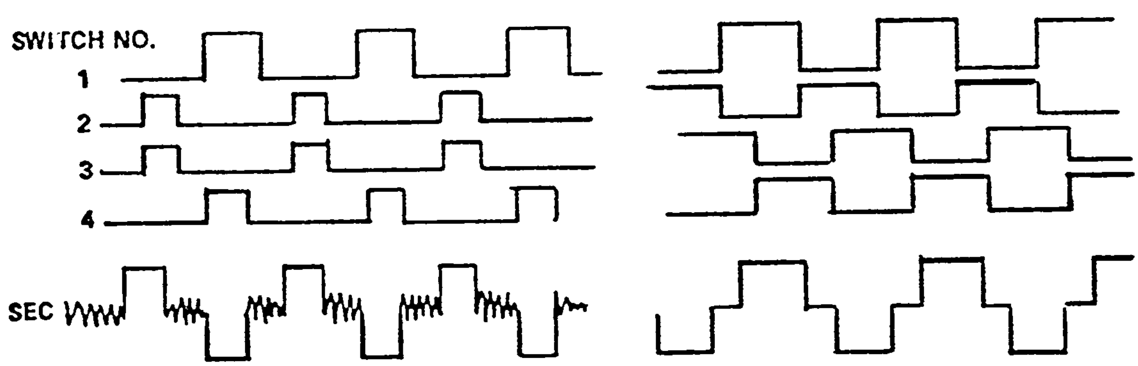


TRIANGLE WAVE GENERATOR
SQUARE WAVE GENERATOR
CONTROL CIRCUIT
+12VDC
.01 uf
100PF
LM 201
OR
LM301
A6
A5
A7
AMP
5.1K
FREQ
20K
12K
100PF
1M
1K
10K
0V
-12

D-706048



a)



b)

Figure III-W

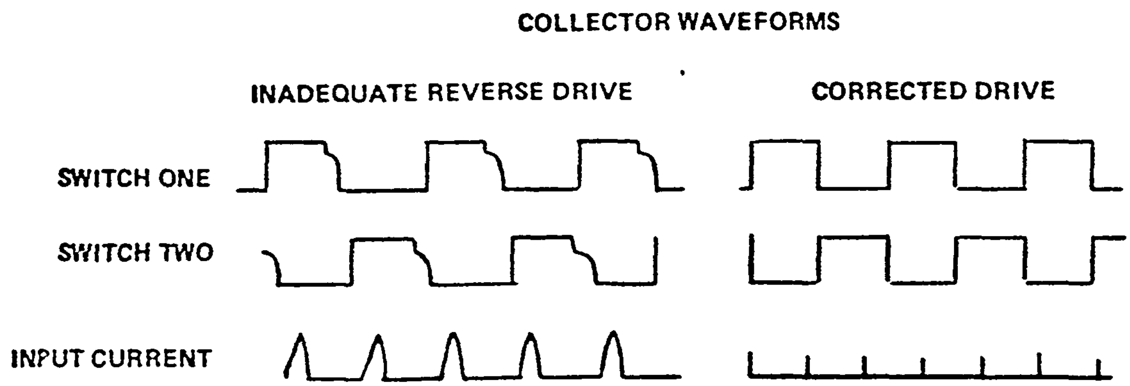


Figure III-X

As shown by the figure above, the finite response time of the transistors still results in a momentary short circuit condition. Even with switch overlap reduced to one microsecond, this current is sufficient to cause circuit destruction. As a further means of avoiding this condition, a delay was introduced in the control circuit to delay, as shown in Figure III-Y.

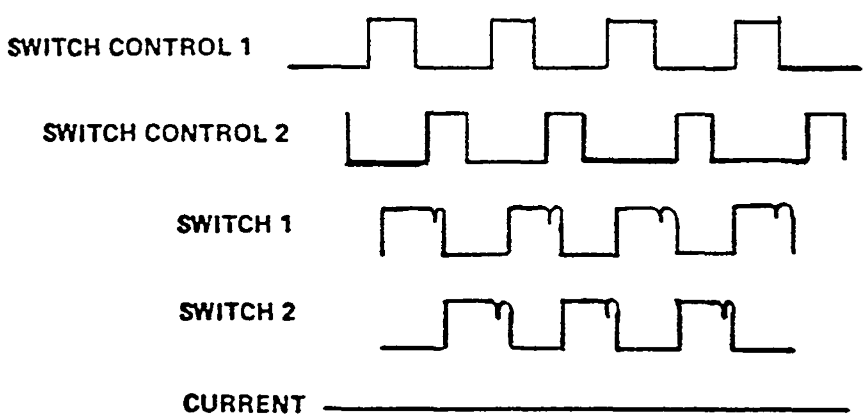


Figure III-Y

E) Phasing Considerations (Continued)

As shown, the assymmetrical control signal compensates for switching delays, completely eliminating the undesired overlap. This delay was ultimately incorporated into the switch base drive circuit.

Buffer Drivers

The function of the buffer drivers is to convert the low level control signal from the logic to a larger signal having sufficient current capacity to drive the switch banks. To accomplish this, two medium-power transistors are connected in a Darlington configuration operating from +24 volts. The output is transformer coupled to the base driver circuits. Ferrite cores are used for the driver transformers.

Feedback Circuit

The feedback circuit performs several functions:

Senses the output current and maintains it at the selected level.
a. Provides current limiting during output short circuit conditions.
c. Compensates for input voltage fluctuations.

Current sensing is accomplished by inserting a fixed resistor of k value in series with the output circuit. This shunt resistor is precisely calibrated to provide a known voltage drop for a given current through the resistor. A low-level voltage is therefore generated, which is proportional to the output current. This voltage is applied to an operational amplifier where it is compared with a level established by the current select switch. Any difference in levels results in an error output from the operational amplifier, which is proportional to the amount of current deviation. This output, applied to the correct terminal of the delay multivibrator in the control circuit, results in a shift in switch phasing which restores the current to its correct level.

Since illtering is required on the operational amplifier input, the response of the circuit is not instantaneous. Current transients of less than several cycles duration will therefore not be corrected by this circuit and must be handled by other means. This delay can be made adjustable so as to allow the operator to select the response rate most suited to this operation.

The foot pedal is so connected as to allow the operator to vary the arc current between zero and the maximum as established by the selector switch.

F) Transformer Design

The transformer design is the heart of the weight reduction solution The core area, windings, exitation and frequency are related in the following formula (square wave exitation):

$$10^8 V = 4NABrf$$

where V = peak value of the square wave (volts)
N = number of primary turns
A = cross sectional core area (CM^2)
B = core flux density (gauss)
f = frequency of applied square wave (Hz)

The formula can be rewritten as:

$$NA = \frac{10^8 v}{4Bf}$$

The product NA is representative of transformer weight copper (N as turns) and core (A as area). Since V is determined by 440v line (650 VDC rectified) and B is limited by the saturation flux of the core material, the frequency must be increased in order to reduce transformer weight. Grest care must be taken in high frequency transformers to minimize core loss since core losses are roughly proportional to the square of frequency.

The first transformer was wound on a commercially available Arnold Engineering core It was the only commercial core large enough (3 1/2" ID). The core is permalloy type core containing molybendum and nickle along with ferrous material.

First tests on the transformer were unsuccessful. The high nickle content causes a very low permeability in the core material; this makes it unsuitable for transformer applications.

The next transformer was wound on a specially made core obtained from Ceramic Magnetics of New Jersey. The dimensions were 5" ID 7" OD, 1" high (this is a core area of 1 sq. in. or 6.45 CM^2). The transformer was wound in house; the windings were somewhat loose, and considerable leakage inductance was observed. There were 175 primary turns and 11 secondary turns wound bifilar.

First tests *on* the second core were satisfactory. The transformer efficiently transmitted. The power, however, due to the excessive leakage inductance, semiconductor damage was observed, plus regulation was poor; the transformer exhibited excessive "drooping" characteristics. Steps were taken to minimize the inductance.

F) <u>Transformer Design</u> (Continued)

A smaller core was ordered from ceramic magnetics. The dimensions were 4 1/2" ID 6" OD and 3/4" high. (This gives a core area of 3.63 CM^2.) The transformer was wound by Magnetic Devices of San Diego. The engineers at Magnetic Devices suggested a winding technique to minimize leakage inductance. The primary and secondary copper areas were divided in three. Layers of primary and secondary windings were intermeshed (first a primary winding, then a secondary, then a primary and soon for three layers). The three secondary windings were paralleled, as well as the primary windings. This technique is illustrated schematically in Figure III-Z.

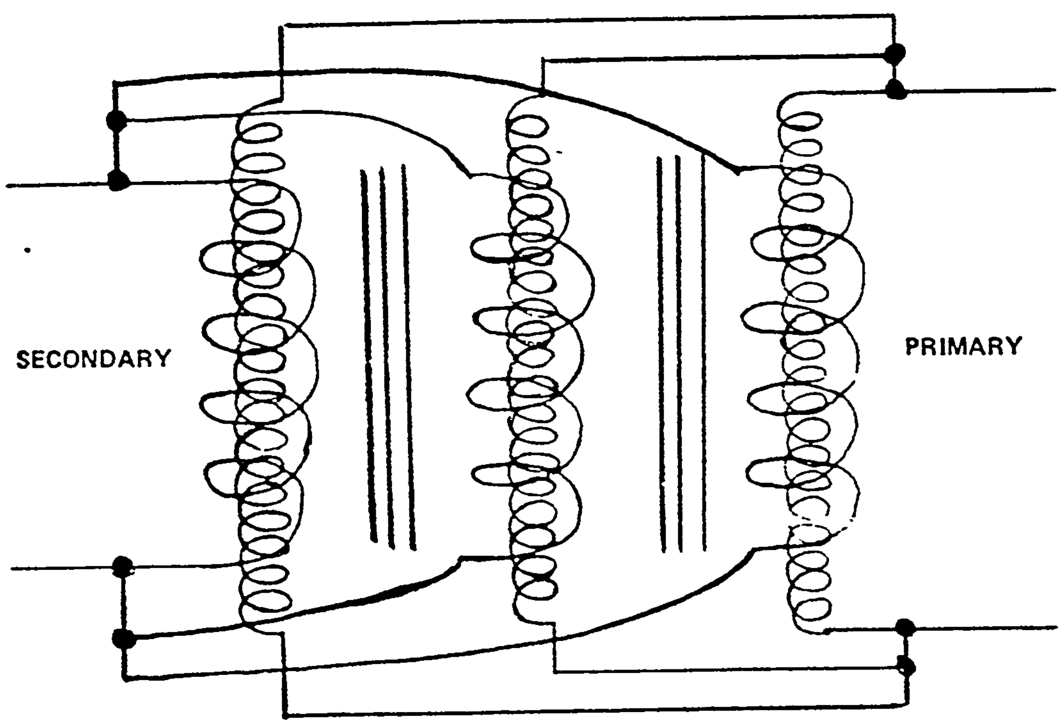


Figure III-Z

F) Transformer Desires (Continued)

The effect of the tight windings and the intermeshed primary and secondary was approximately a 70% reduction in the leakage inductance. The transformer then efficiently transmitted the energy from primary to secondary with a considerably reduced "drooping" characteristic.

The material used in the Magnetic Ceramics cores was called MN60, which is a manganese and ferrite alloy. The core had high flux density value (5600 gauss) and high permeability (about 10,000). These characteristics, in conjunction with low core and hysteresis loss, *make* this an excellent high frequency power transformer material.

G) Mechanical Packaging

Minimum size and weight were the primary considerations in packaging this power supply. Heat sink selection and a packaging concept were the major considerations. Heat dissipation was investigated first. The major heat producing components producing 85% of the heat dissipation, were the (1) input bridge, (2) switching circuits and (3) output bridge. The basic heat sink design and cooling requirements were designed around these components. The remaining components with relatively small power dissipation were to be mounted and heat sinked internally to the case.

Since as much surface area as possible would need to be reserved for heat sinks, the external interface components, such as switches, indicators, the meter and output connection, were located on the front panel. Possible stacking of power supply units for parallel operation was also kept in mind for future units. This consideration along with the fact that there will be dead air space on the bottom of the unit, made it undesirable to consider the top and bottom for heat sinking. This left the surface area of three sides for cooling.

Heat sinking to be used must be one sided or else arranged in a manner so as to keep all wiring internal to the case and provide for ease of sealing and exposed electrical conductors or connectors. All external heat sinking will be connected to a common ground **to** eliminate electrical shock hazard. Electronic components must be insulated from the heat sinks, thus increasing thermal resistance between component case and heat sink. Considerations such as thes must be made in the thermal analysis.

A survey of available heat sinks was made and literature was obtained from the following manufacturer:

Republic - TOR	Gardena, California
Thermalloy Inc.	Dallas, Texas
International Rectifier	El Segundo, California
Wakefield Engr. Inc.	Wakefield, Massachusetts
Astrodyne Inc.	Wilmington, Massachusetts
Delbert Blinn Co.	Pomona, California
IERC	Burbank, California

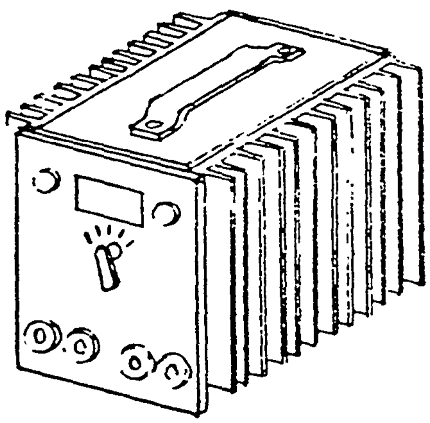
G) Mechanical Packaging (Continued)

The majority of available heat sinks are extrusions, which can be combined to suit the component sizes and requirements. Several approaches were considered for cases such as 1) thin walled aluminum casting with cooling fins, 2) a combination of aluminum heat sinks extrusions bolted to a skeleton frame, or 3) possibly a combination of aluminum (extrusions) welded together to form a one-piece box structure. Also to be considered would be special extrusions made to suit this application. Because of cost, availability and ease of disassembly, it was decided to select standard heat sinks and bolt them to a frame structure for the first packaged breadboard unit. This would help minimize the time to assemble the breadboard for test at possibly some sacrifice in weight and size. In further refinements, a more compact, lighter weight design could be achieved. On the first breadboard, water tight sealing was not imposed. However, all exposed components were to be covered and insulated, with the result being a completely enclosed unit.

The transformer, because of its *size* and weight, is a primary consideration of the components to be mounted internally. It must be firmly mounted to the structure for ruggedness and conveniently mounted for making electrical connections. It must be low and centrally mounted because of its effect on the center of gravity.

With these considerations, two approaches to packaging were considered for the breadboard.

1) A sheet metal box with heat sink extrusions mounted externally.



Box Mounted Unit

G) Mechanical Packaging (Continued)

Considerations of this type of package:

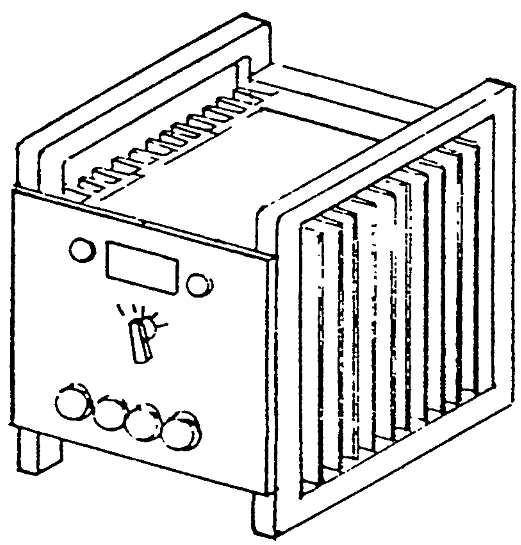
(a) Need rigid box structure.

(b) Handle must be rigidly attached and an integral part of the structure.

(c) Sharp corners of extrusions are a problem and must be well rounded or protected so as not to be a hazard.

(d) Front and top panels and/or extrusions must be removable for access to internal circuitry.

2) A tubular, external structure which would provide rigidity with heat sinks mounted to it or to a lightweight inner case.



Frame Mounted Unit

G) Mechanical Packaging (Continued)

Considerations for this type of package:

(a) Frame would provide rigid protective structure.

(b) Handles to be a part of the frame.

(c) Tubular structure to protect thin fins of heat sinks from damage due to handling, and also provides protection for sharp corners of extrusions.

(d) Provides a structure which can be easily adapted to stacking.

(e) Internal case would be lightweight and of minimum structure.

Of these two approaches, the latter appears to be the most favorable. Once heat sinks are selected, a final determination can be made. In either case, extrusions will be mounted vertically for maximum efficiency and some air space will be provided beneath and above each heat sink to improve air flow.

H) <u>Evaluation</u>

During the investigative portion of this project, extensive tests of sub-assemblies were made, comparisons were established, and designs were chosen. This period corresponds to points one, two and part of three of the Plan outlined in the Objectives Section of this report.

The input rectifier was chosen as the originally designed three-phase bridge rectifier. There were virtually no other choices for this function. The transistors were chosen for the switching circuitry as being the best means of achieving success (rather than SCR's) mainly for their control and advantages (Figure III-R provides a detailed comparison). The digital control circuitry was chosen for its simplicity and accuracy, rather than the analog control. The transformer design was optimized during this period as far as leakage inductance and efficiency are concerned. Mechanical packaging and heat sinking of the electronic components was investigated for feasibility. Commercial manufacturers of heat sinks were identified and analysis of their products was performed. A rough conception of the mechanical packaging was established.

IV Finalizing the Candidate Design

A) Further Development

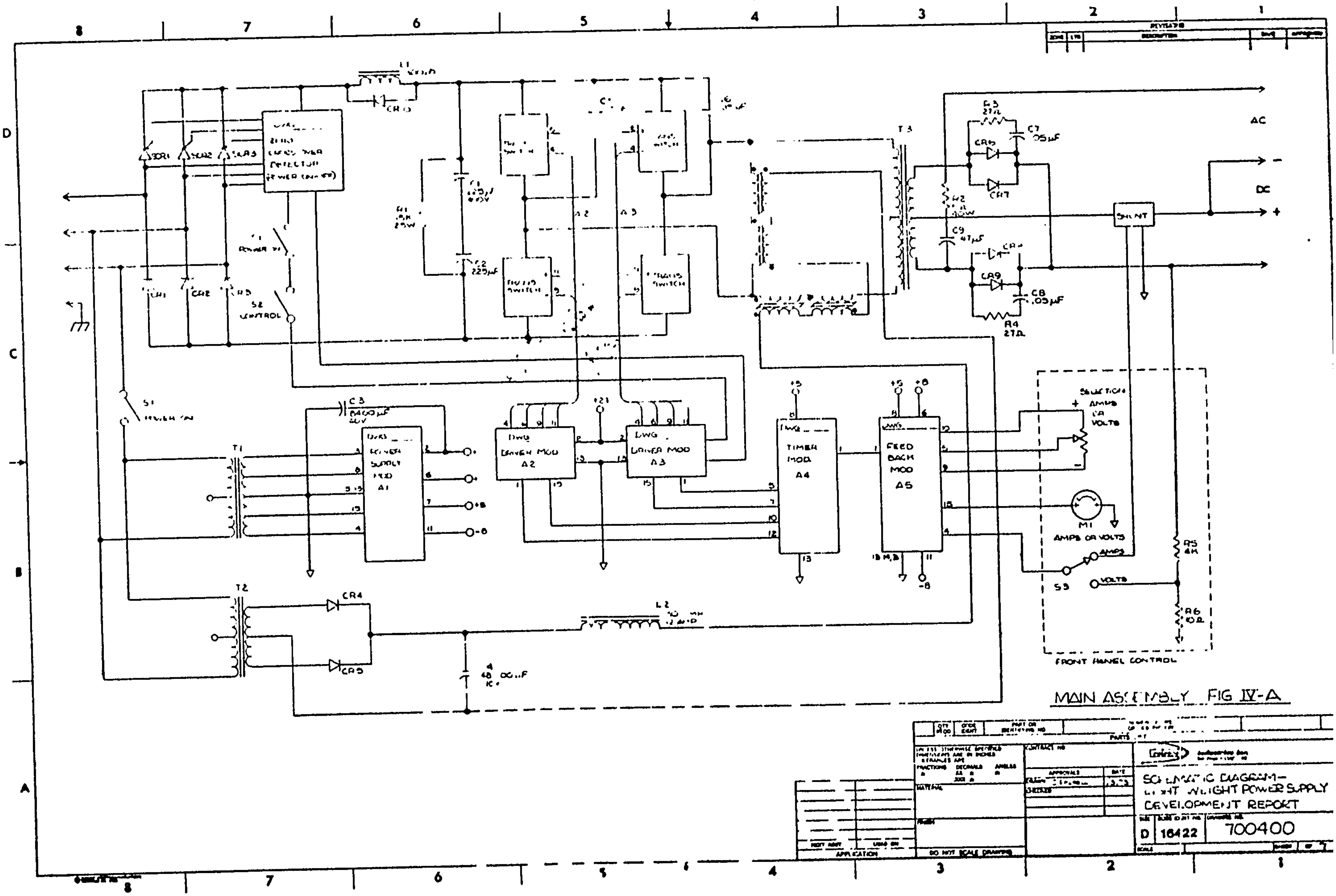
As indicated in the previous section, the transistor switching circuitry was selected as the most promising design for achieving a demonstrable unit within the time and budget limitations. This decision was based on the initial research outlined in the previous section. Full development of the transistor switching circuit was required in order to have an operational system.

Figure HI-R lists some of the inherent problems of transistors for this switching application; current and voltage ratings being the most limiting and severe problems, and also charge storage phenomenon causing switching efficiency problems. Forty Delco DTS 804 transistors were arranged in a bridge configuration as shown in Figure III-W of the previous section.

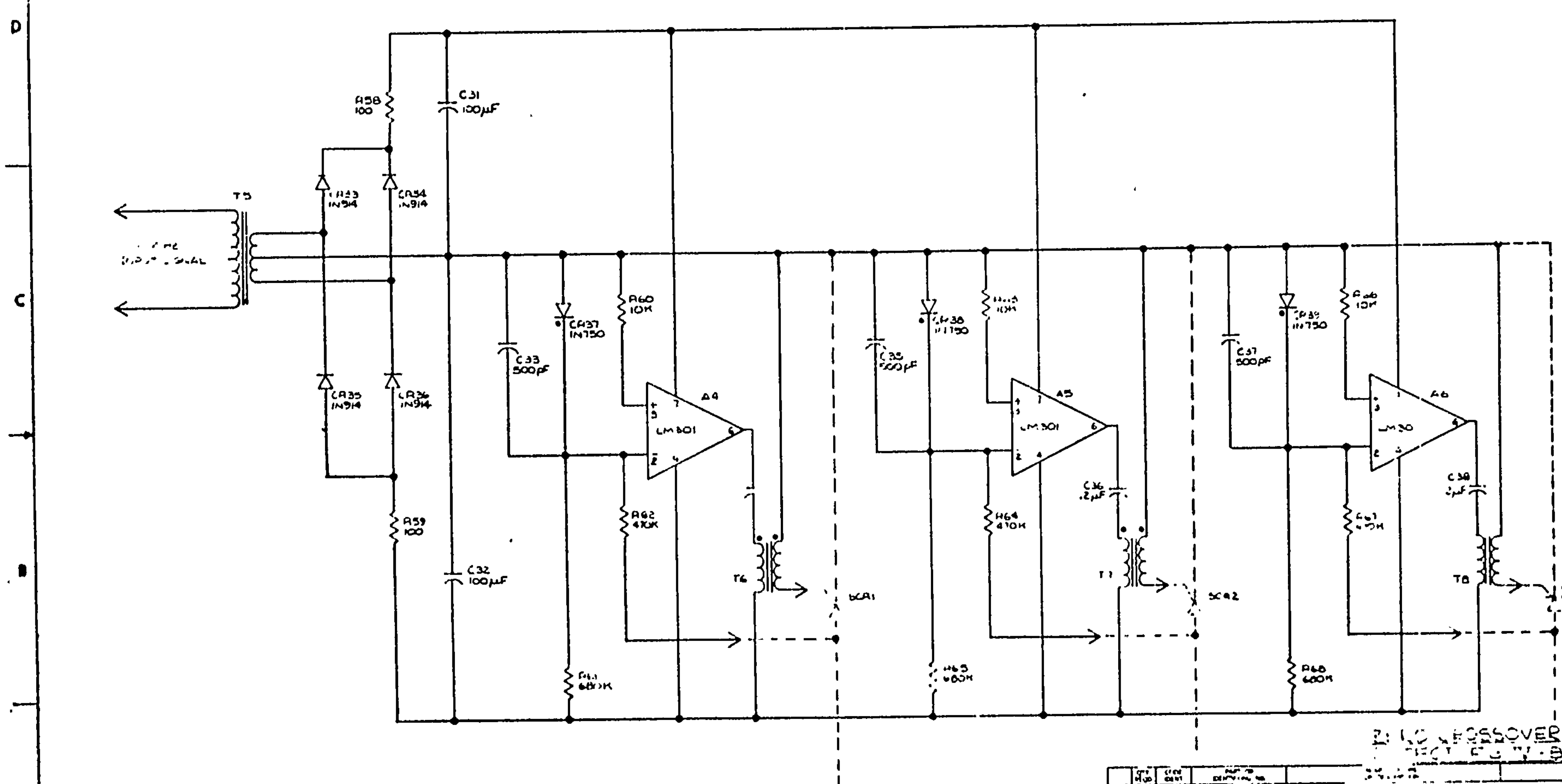
First tests were promising using the candidate design. welding was conducted using TIG and SMAW. However, random failures occurred, and significant switch heating was observed. The failures were attributed to voltage and current transients occurring due to the time delay of the control circuitry. Since the feedback circuit is basically an integrator reacting to changes in average value of the current, the control circuit could not react instantaneously to short circuits. During this period of control recovery, excessive currents and voltages appeared at the transistor, resulting in their destruction. It was decided, therefore, that a subcycle current limiting would have to be incorporated independent of the control circuitry.

Resistive current limiters were impractical due to the excessive dissipation they would cause. Therefore, reactive current limiters were used in the form of linear inductors and linear capacitors. The inductive current limiting was successful in reducing the semiconductor failure. However, it had some undesirable characteristics. The inductor was placed in the primary line, as shown in Figure IV-H. A 442μh inductor provides a 20 r reactance to the fundamental 7.2 K Hz frequency, thus limiting the fundamental primary current to 30 amps peak.

Due to the slight nonlinearities in the inductor, and the fact that significantly lower power levels in the 3rd, 5th and 7th harmonics were passed, the inductor method of current limiting provided a V-I characteristic shown in Figure IV-1 that would be undesirable for welding applications. As can be seen, at an arc potential of 23V, only 150 amps could be delivered (point 1). Point 2, 40 volts at 300a is far beyond the capability of the unit. Also, the subcycle current waveform contained high peaks at the end of the cycle, as shown in Figure IV-J. This is typical of inductive loads.

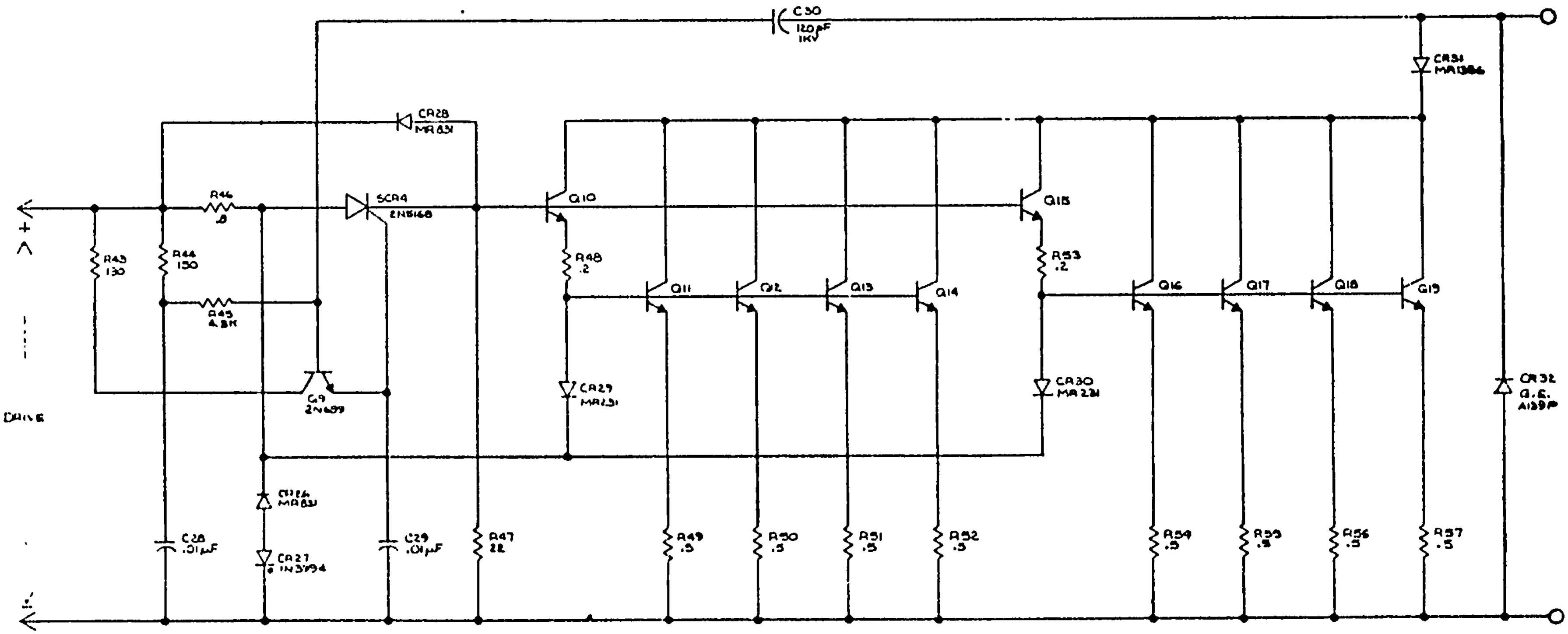


MAIN ASSEMBLY FIG IV-A



SCHEMATIC DIAGRAM – LIGHT WEIGHT POWER SUPP DEVELOPMENT REPORT

D 18422 100 400



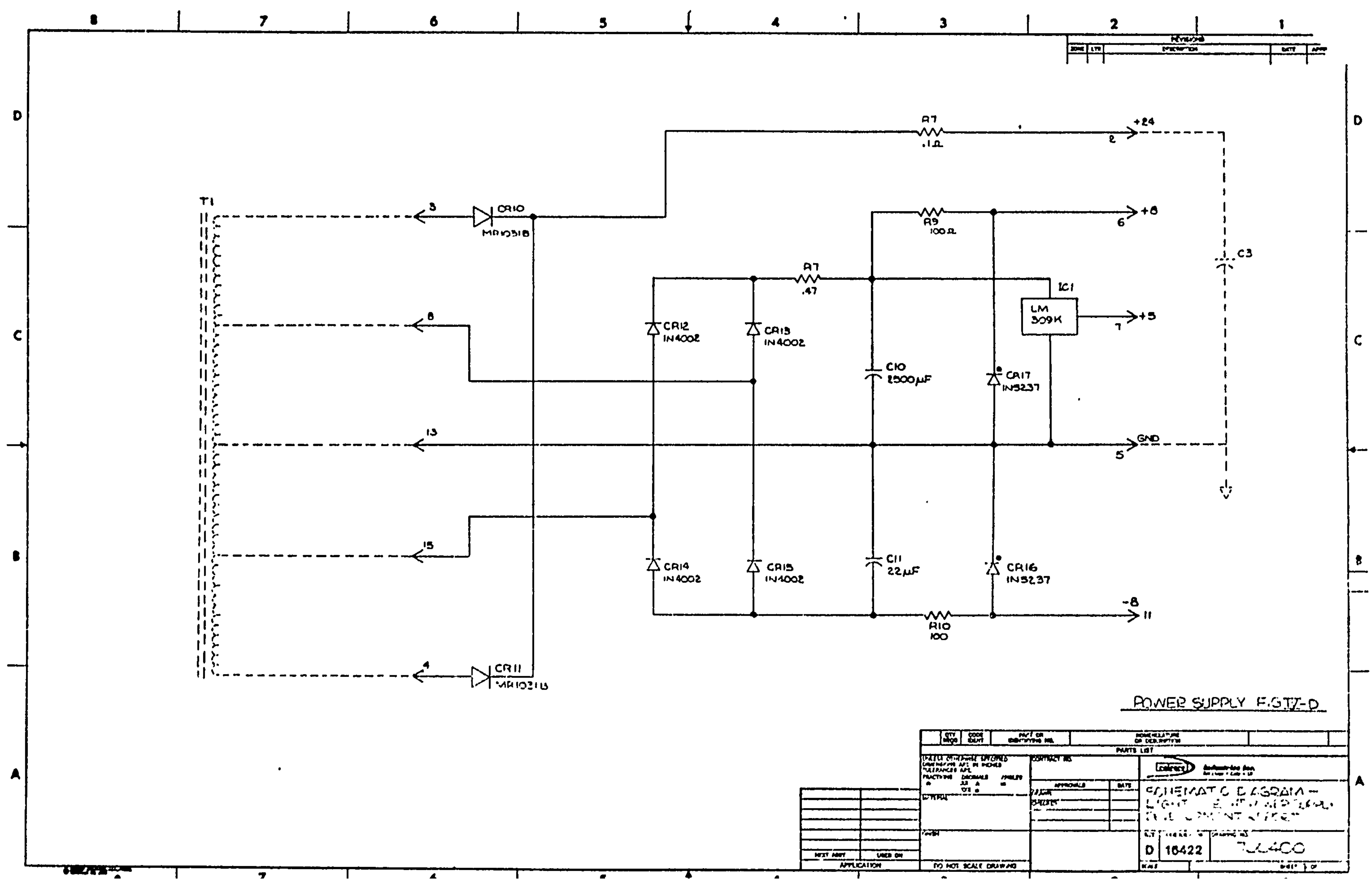
Q10 THRU Q19 = DTS 804 (DELCO)

TRANSISTOR SWITCH FIG IV-C

SCHEMATIC DIAGRAM - LIGHT WEIGHT POWER SUPPLY DEVELOPMENT REPORT

D 18422 700400

DO NOT SCALE DRAWING


T1
3
CR10
MR1031B
R7
.1Ω
+24
2
C3
R9
100Ω
+8
6
R7
.47
IC1
LM
309K
+5
7
8
CR12
1N4002
CR13
1N4002
C10
2500μF
CR17
1N5237
13
GND
5
15
CR14
1N4002
CR15
1N4002
C11
22μF
CR16
1N5237
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R10
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4
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PARTS LIST
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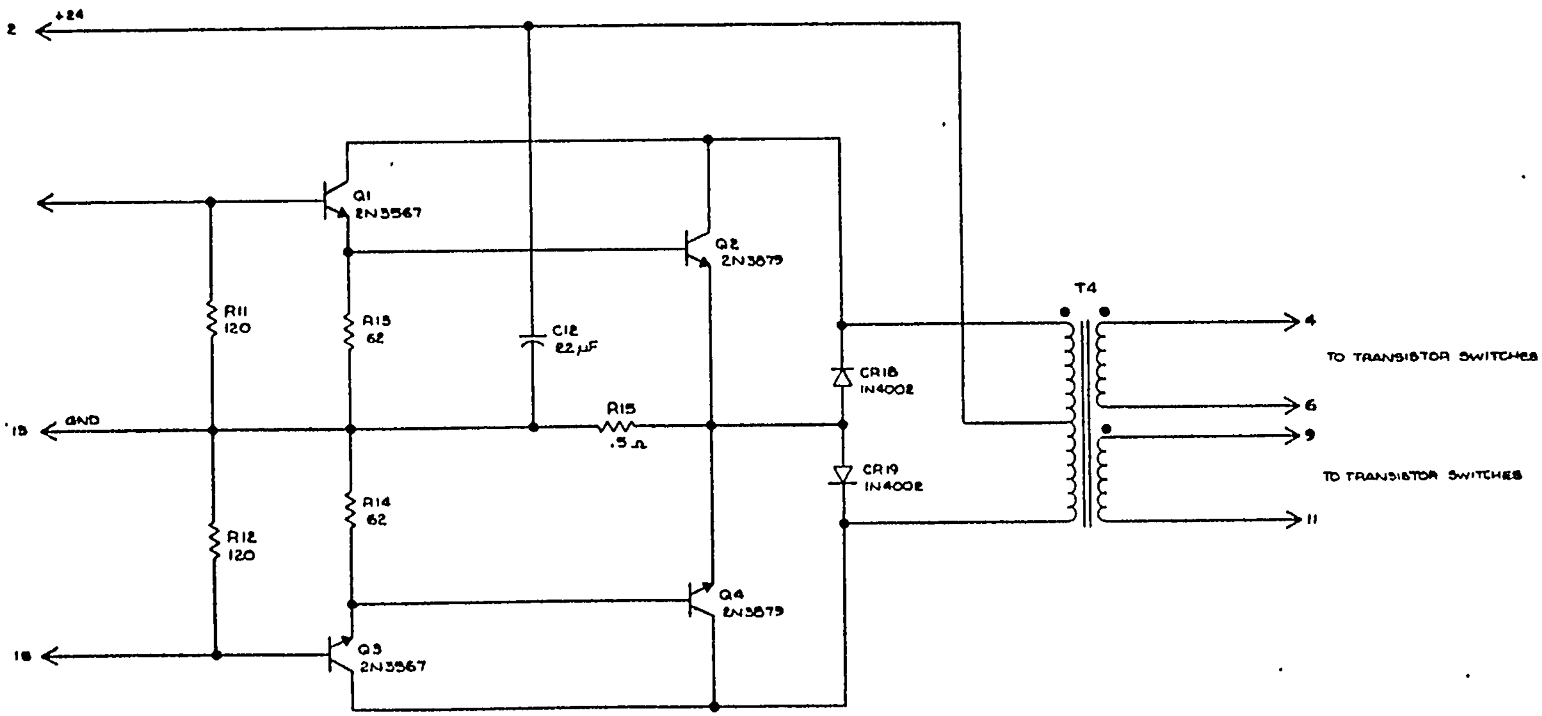
16422



SWITCH DRIVER FIG IV-E

PARTS LIST

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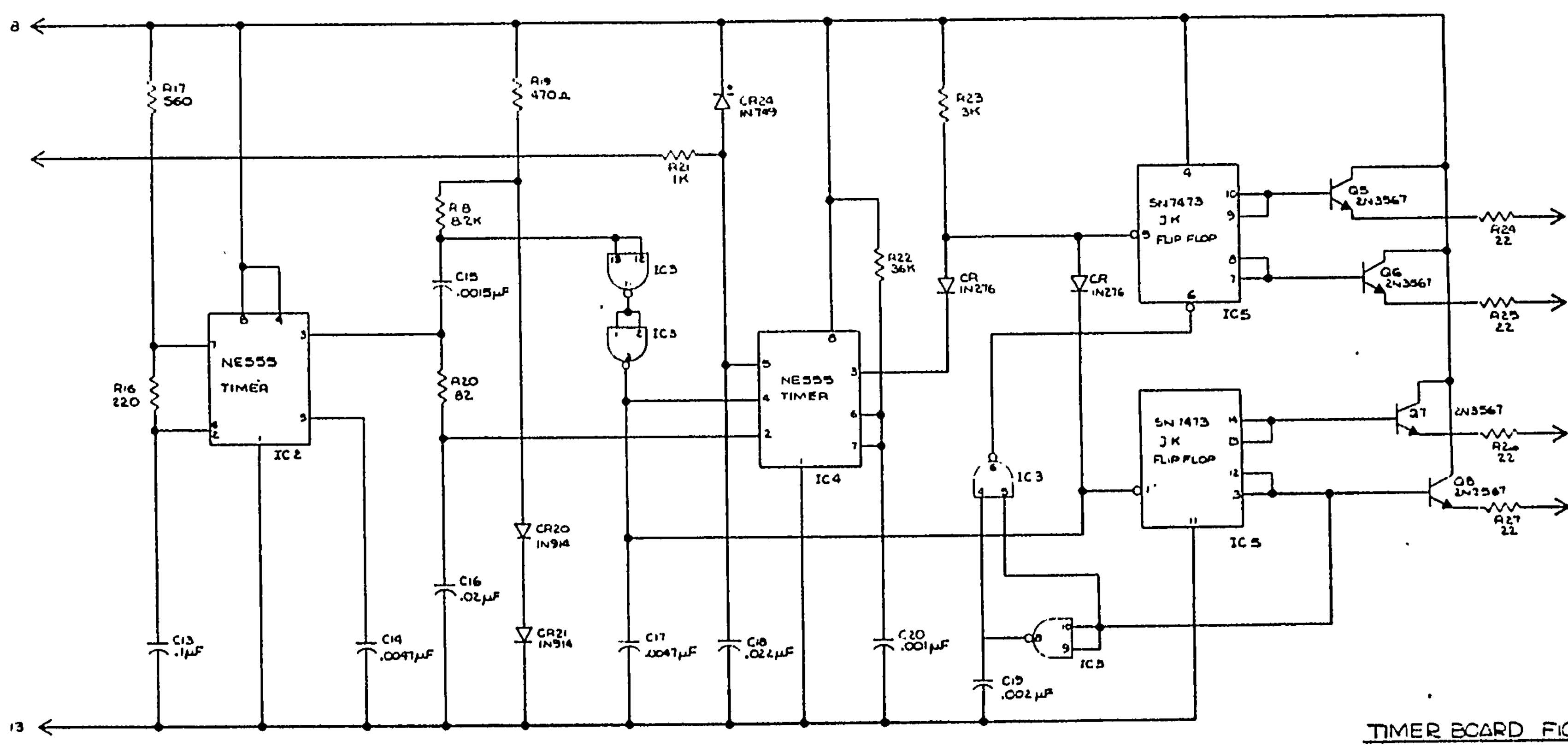


TIMER BOARD FIG IV-F

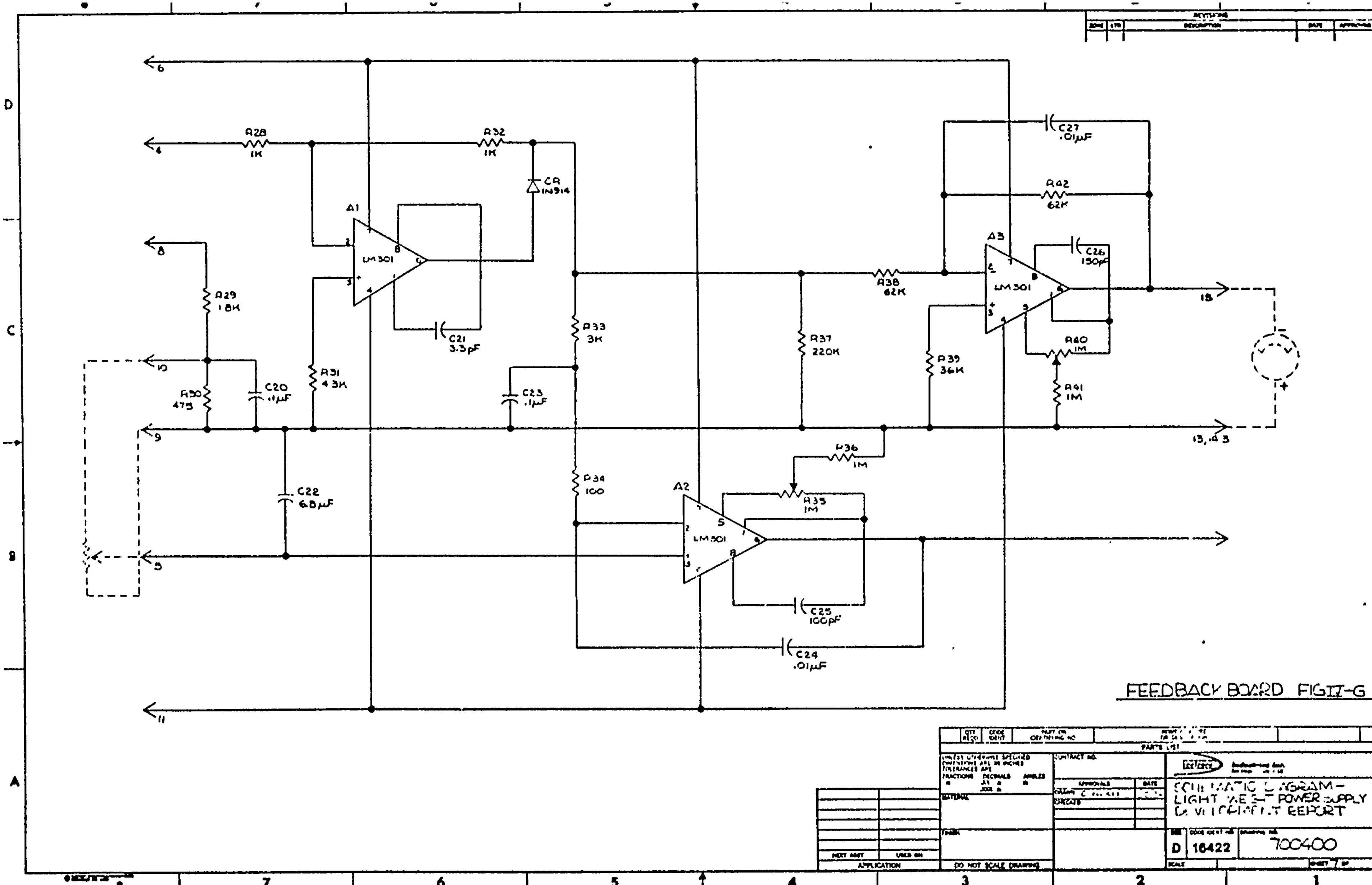
SCHEMATIC DIAGRAM — LIGHT WEIGHT POWER SUPPLY DEVELOPMENT REPORT

SIZE	CODE IDENT NO	DRAWING NO
D	16422	700400

DO NOT SCALE DRAWING

APPLICATION

SHEET 6



FEEDBACK BOARD FIG IV-G

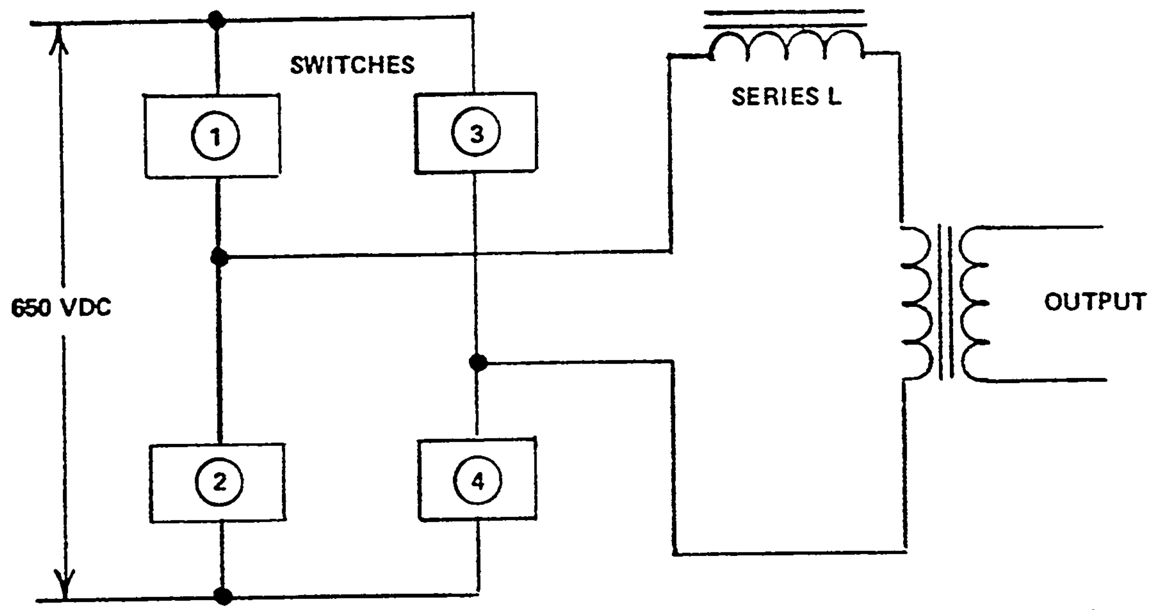


Figure IV-H

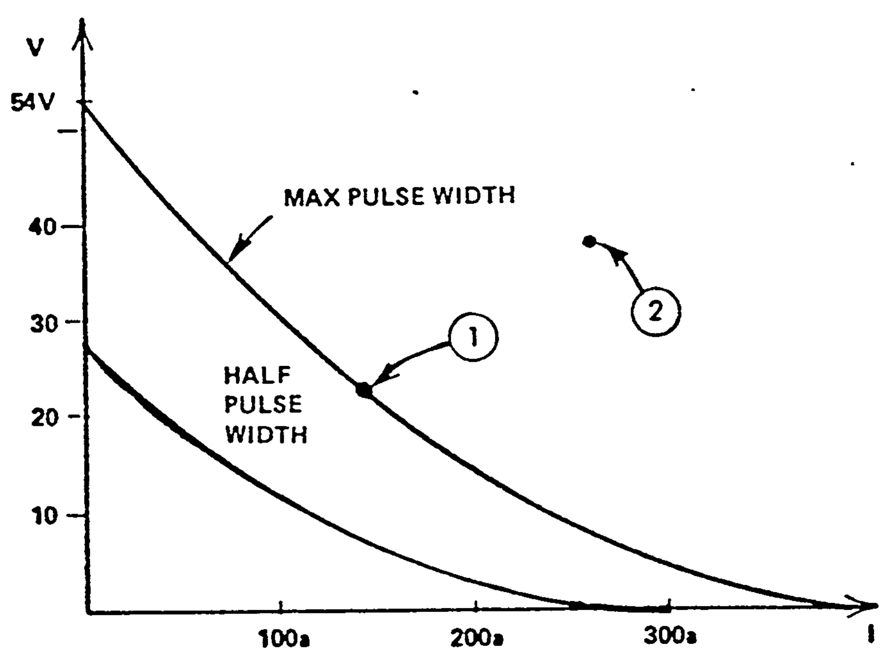


Figure IV-I

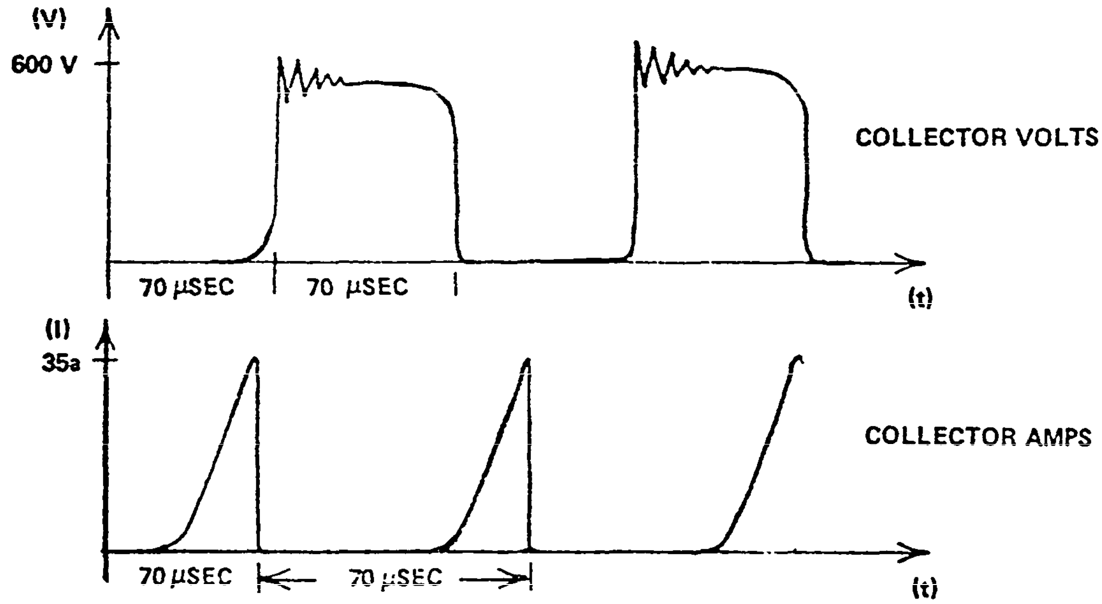


Figure IV-J

A) Further Development (Continued)

Linear capacitor elements were then tried in place of the series inductor in order to provide the same effect, except capacitive reaction was used instead. The capacitor now could effectively limit the fundamental current; however, its limiting effect became negligable at the higher harmonics (of the square wave) due to the fact that capacitive reactance is inversely proportional to frequency. The capacitive limiting reactance also had an undesirable effect of resonating with the leakage inductance of the power transformer. Large current peaks caused by the resonance would result in component destruction.

A) Further Development (Continued)

A saturable reactor current limiting was then used. Like the transformer, the saturable reactor's weight is also considerably reduced by the increased frequency. Electronically it is placed in series with the primary winding in the place of the series inductor of Figure IV-H. The total weight of the reactor circuit was approximately twenty-five pounds. This figure includes the reactors bias inductor and DC bias supply. The circuit is shown schematically in Figure IV-K. Weights are also indicated. These weights can be significantly reduced (to about seventeen pounds) by also incorporating high frequency techniques, rather than conventional 60 Hz techniques.

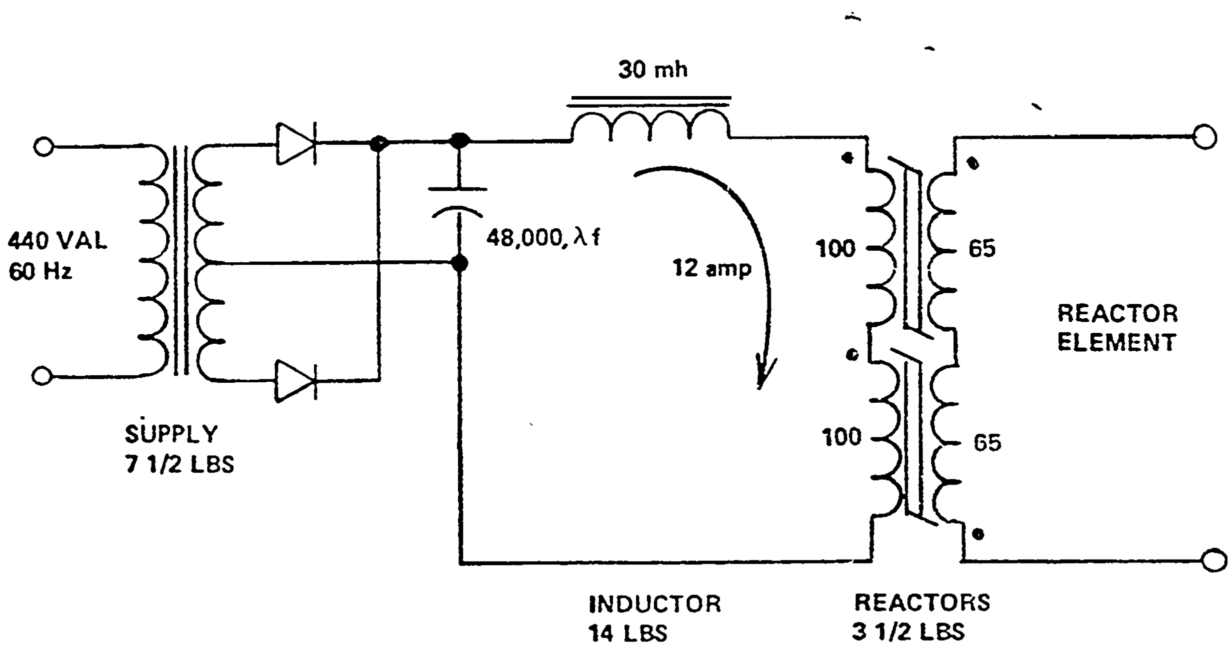


Figure IV-K

A) Further Development (Continued)

Tests on this reactor circuit were very promising. Current limiting was successfully achieved at 300a (secondary current). Welding was conducted at full line input voltage. Semiconductor failures during welding (SMAW) were eliminated. The output V- I characteristics are shown in Figure IF-L. The collector current waveforms are shown in Figure IV-M. These may be compared with Figure IF-I, J for improvement. Output power capabilities were significantly increased with increased reliability.

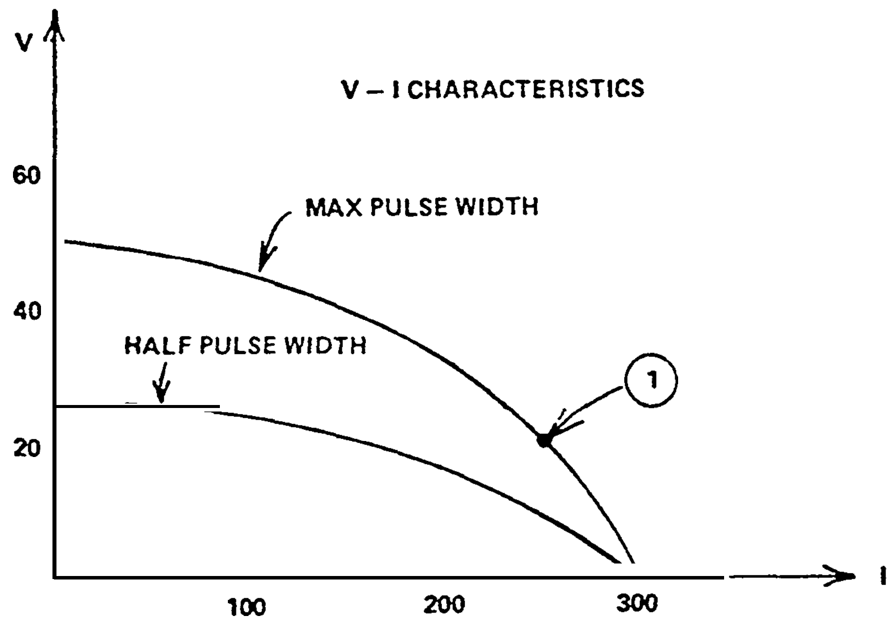


Figure IV-L

Note at point 1, 260 amps are available at 25 volts; compare with Figure IV-1 where only 150 amps are available at 23 volts.

Further tests with the saturable reactor limiting and transistor switching revealed that the reactors caused a turn-off strain on the transistors (voltage transients). Bypass capacitors were added in the circuit as shown in Figure. IV-N; however, in eliminating the turn-off strain due to voltage transients and stored charge in the transistor a turn-on strain in the form of a current peak was cussed by the bypas capacitors. This turn-on strain was eliminated by creating a time delay between the "turn off" of switch 1 and the "turn on" of switch 2 (and visa versa) and by adding a small saturable reactor across the transformer (also shown in Figure IV-N). The small reactor then, its energy stored in the field, was able to provide change for capacito recovery within the the turn-on to turn-off time delay. The turn-on strai was similarly eliminated in switches 3 & 4.

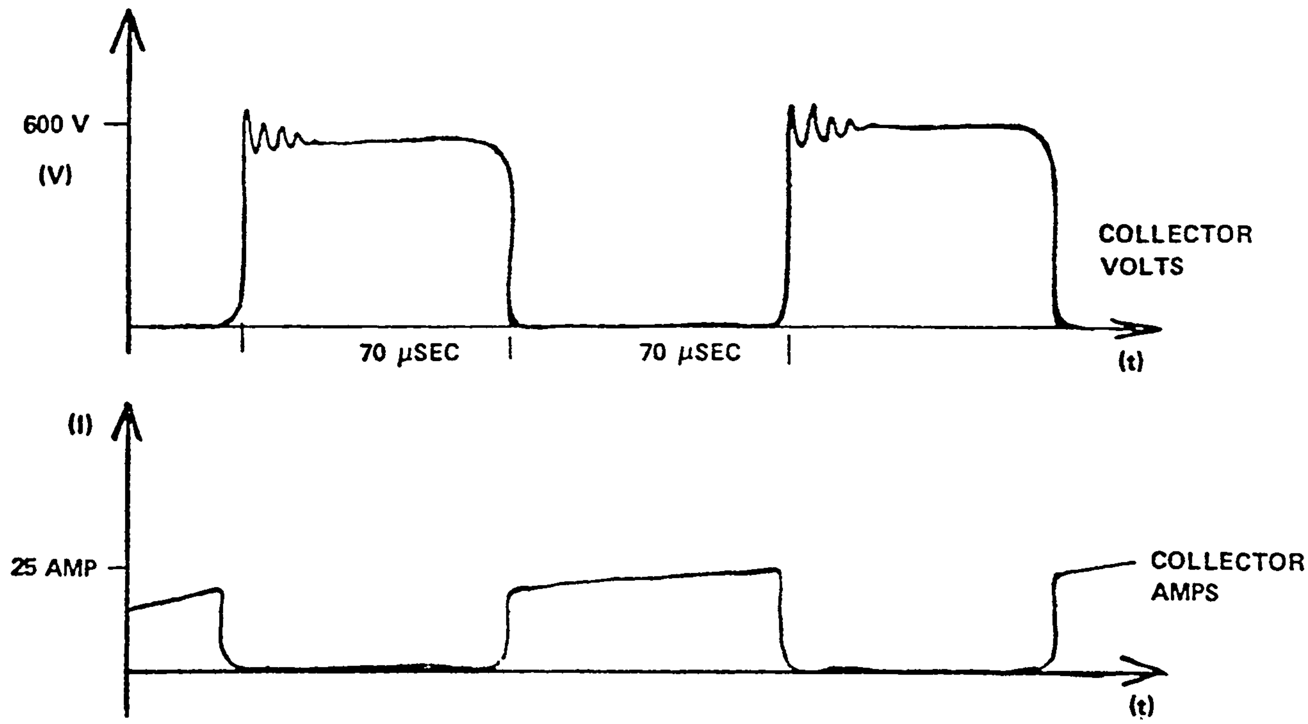


Figure IV-M

Note the even 25 amp current level during the "on" period of the transistor as opposed to the 35 amp peak shown in Figure IV-J.

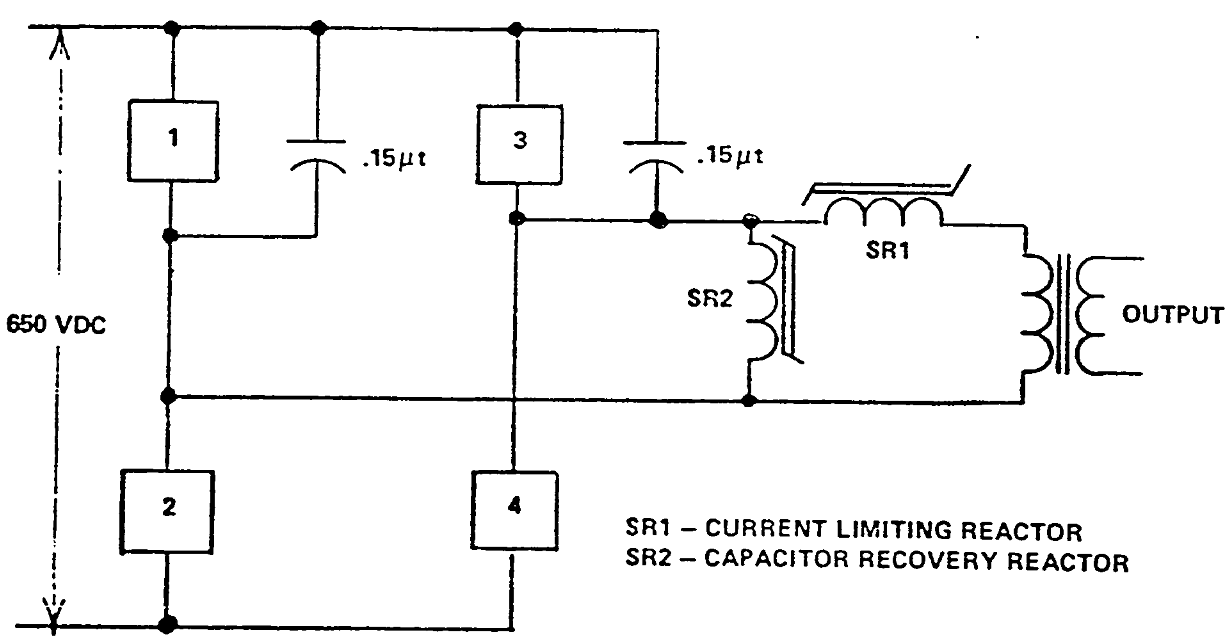


Figure IV-N

A) Further Development (Continued)

This capacitor recovery technique was quite successful. This technique could eliminate the SCR's main problem (Figure III-R) 1 variability, by using a variable limiting reactor (SR1) 2 turn-off, by using the above efficient capacitor recovery technique. The result for the transistor circuit was increased switch efficiency and thus increased realizability

Further tests indicated that transistor failure also occurred during the instant of power initiation. It was discovered that in addition to lack of control circuit stabilization upon power initiation, secondary peaks caused by the inductance of the 440 line charging the input capacitor (650 VDC filter) resulted in voltages in excess of transistor limits. The zero crossover SCR switch shown in Figure IV-0 was used to reduce the secondary peaks. The 650 VDC would rise with the 1/4 cycle of the 60 Hz cycle, thus eliminating the possibility of a 1 or 2 μsec step to full voltage with a plain rectifier and a series switch. Secondary peaks were virtually eliminated. The SCR firing logic also contained a programmed delay to allow for control circuit stabilization.

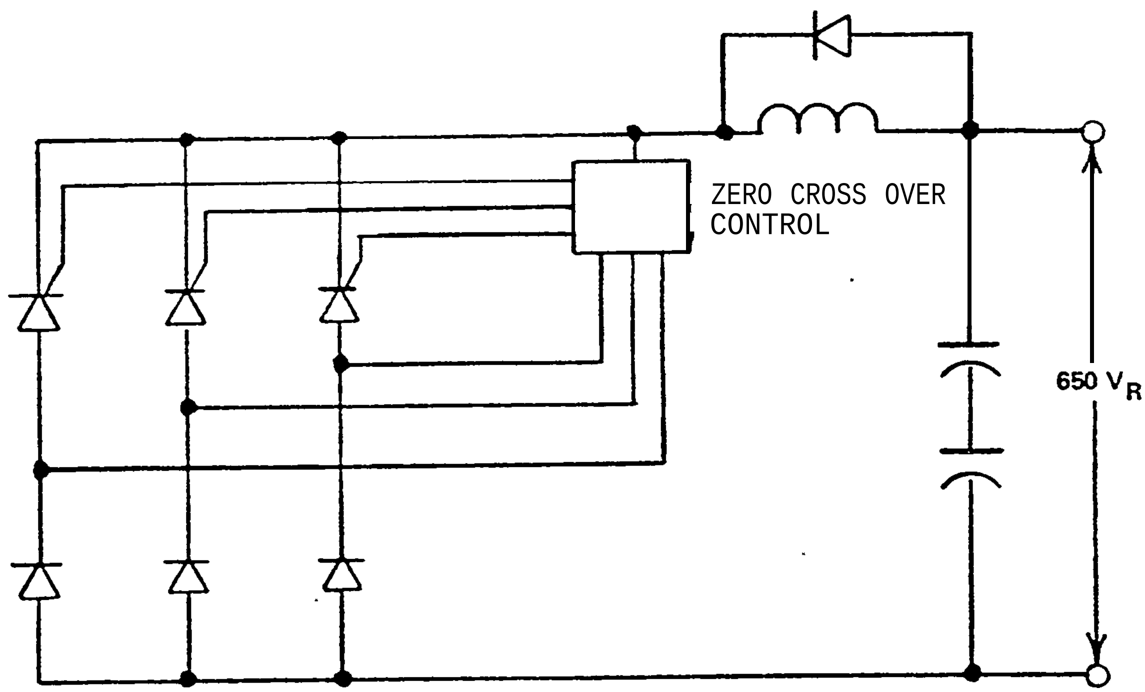


Figure IV-O

A) Further Development (Continued)

During the period of "Further Development of the Transistor Switches", the following transistor problems were eliminated:

1) Current peaks were eliminated by effective saturable reactor current limiting. Power output and realizability were also increased.

2) Heating due to stored charge was reduced by bypass capacitors and reactor recovery for the capacitor. Reliability and efficiency were also increased.

3) SCR contactor was incorporated to reduce transients and provide for smooth turn-on.

B) Electronic Circuit Diagrams

See Figures III-S, T, U and V.

The following paragraphs of this section are devoted to the documentation and explanation of the unit resulting from this research. The circuit diagrams are first followed by theory of operations and design calculations.

c) Theory of Operations

Refer to drawing Number 1000, Figure IV - A

This is a schematic of the main assembly. The basic diagram shown in Figure IV-4 can be visualized in this main assembly drawing. The input rectifier consists of SCR 1, SCR 2, SCR 3 and D 1 to D 3. Next comes the filter L1, Cl and C2; then the transistor switch bridge configuration. The transformer network consists of SR 1, SR 2 and TR 3; the output rectifier consists of D 6 to D 9. The feedback loop can be seen starting with reference signal at RRl and current feedback module A 5 compares the reference and controlled signs. giving an error signal to the Timer Module A4. The Timer module then controls the driver modules A 2 and A 3 which control the transistor switches 1, 2 , 3 and 4 SCR 1 to SCR 3 perform the added function of an on/off switch, a lightweight version of a contactor. The SCR's are controlled by the zero-crossover detector (DWG No. 1001) and S1 (power on) and S2 (contactor). TR1 and the power supply module Al derive and regulate the logic supplies from the 440 VAC 60 Hz line. TR2, D4, D5, C-1 and L2 provide the DC bias to the saturable reactors SR1 and SR2 (12 amps).

D) Design Calculations - Some of the major design calculations are presented in order to give further insight into the electronic operation of the system. Refer to Drawing No. 1000, Figure IV-A.

D) Design Calculations (Continued)

Switch Bypass Capacitors

Since the stored change in DTS 804 cannot be swept out faster than about 1.5 sec., it is desired to limit the voltage rise to no more than 20% in 1.5 sec. or 130 V in 1.5 sec. Now since I primary is limited to 25 amps (by SR2), then Ip max = 25 amps.

$$I = C \frac{dv}{dt}$$

$$25a = C \frac{130}{1.5} \text{ sec.}$$

$$C = .25\mu f$$

$.15\mu f$ was available at the necessary WVDC

$$C_5 = C_6 = .15\mu f$$

Reactors

SR1 and SR use the same cores cross section of .60 CM $B_s = 10,000$ gauss. Both can have a worst case blocking voltage of the full 650 volts for 1/2 period or approcimately 70 sec.

primary turns

$$Np = \frac{10^8 \quad V}{4B \quad f} \qquad f = \frac{1}{1/2\ T_p} = \frac{1}{70\mu\text{sec.}}$$

$$Np = \frac{10^8 \quad .650}{4.20,000} \ \frac{1}{70} \qquad B = 2B_s \text{ (from full negative to full positive saturation)}$$

$$Np = 56.98 \text{ turns}$$

65 turns were used

$$N\ sec = \frac{\text{Ip Max}}{\text{I Bias}} \quad Np = \frac{25}{15} \quad .65 = 108$$

D) Design Calculations (Continued)

Use N sec = 100 turns for SR2. Now SR1 is used for recovery of C5 and C6 allowing 60μsec. out of 140μsec. for worst case recovery (30μ sec. each polarity).

$$ip = c \frac{dv}{dt} = .15 \frac{650V}{30} = 3.25 \text{ amp}$$

$$N \text{ sec} = \frac{Ip}{I \text{ sec.}} \times 65 = \frac{3.25}{15} .65 = 15 \text{ turns}$$

SR1 N sec = 14 turns

DC Bias for Reactors

C4 - allow a 25% discharge of C4 from Vp. Imperically it was determined that 6.0 volts are needed to force 15 amps through reactor bias windings (allowing for .15r dc resistance for L2). Therefore Vp = 6.0V. V discharge = 25% 6.0
= 1.5 volts

$$i = c \frac{dv}{dt}$$

15a = c 1.5 volts/8.3 ms

C4 = 83,000μf

only 48,000 f was available

C4 = 48,000 f

This value causes approximately a 40% discharge

L2 - voltage variation is approximately $\frac{N \text{ sec}}{Np} \times 650$ using reactor windings

$$V \text{ peak} = \frac{14}{65} .650 + \frac{100}{65} .650 = 1140V$$

Because the two cores are used, the frequency appears at twice primary frequency, about 14 KC or about a 35 sec. period.

$$V = L \frac{di}{dt}$$

now allowing no more than 10% variation in bias current

$$\frac{di}{dt} = \frac{1.5 \text{ amps}}{35 \text{ sec.}}$$

D) Design Calculations (Continued)

$$1140 = L^1 \frac{1.5}{3.5} \cdot 10^{+6}$$

$$L\ 28\ mh$$

use 30 mh

Zero Crossover Detector

Refer to Drawing No. 1001, Figure IV-B. The primary function of the module is to supply the gate pulses in proper time sequence to the SCR contactor (SCR1 to SCR3, Drawing No. 1001, Figure IV-A). The zero crossover detector consists of a power supply (JR5, D33 to D36, C31, C32) and three similar SCR driver circuits, one for each SCR.

Basically, the SCR drivers must sense the positive leading rise of the 60 Hz 440 line sine wave, and fire the SCR as the sine wave *crosse* through the zero volt level. This permits the 650 VDC supply to come up "gradually" in a quarter cycle of 60Hz instead of instantaneously, which causes destructive peaks discussed in part A of this section. A4 and LM1301 is used as a comparator. Consider the SCR 1 first being reverse biased by some negative voltage. A4 is then saturated at the positive supply, the capacitor C34 then charges up to the full plus and minus supply through A-1 and TR6. D37 is a 4.7 V zener diode protecting A4 against arbitrarily large negative voltages on the input R62. (This voltage can be as large as 650V.) C33 is a noise supressor whose time constant (with R61, R62 and D37) is negligible as far as circuit operation is concerned. The voltage on the anode of SCR1 then rises to a positive value as the line voltage crosses zero. When the voltage on the anode reaches +12V to insure anode current, the junctio of R61 and R02 reaches common (zero potential) and A4 switches to the saturated state at the negative supply level. This causes a negative pulse on the dotted winding of TRG due to the charge on C34. This supplies a positive pulse to the gate of SCR1 C34 then rapidly discharges.

The power supply is a critical part of the circuit operation. The input power comes from the 7000 Hz of a switch driver (A3 Figure IV-A, Dwg. 1000) via the power switch and the contactor switch. The supply has a 25 ms rise, due to the time constant of R59, C32 and R58, C31. Therefore, if any SCR has a positive voltage across it (blocking) when the switch is turned on, it will not fire because C34 cannot pass such a gradual slope. The SCR will not fire until the anode potential crosses through zero in a positive direction. Thus, the input circuitry will never be subjected to an input step voltage causing destructive secondary peaks.

Design Calculation Zero Crossover Detector

First of all, it is important to point out why a simple resistor connected anode to gate as follows won't work:

D) Design Calculations (Continued)

Design Calculation Zero Crossover Detector (Continued)

First of all, the resistor does not truly detect a zero crossover; it detects only a positive level; this scheme can produce step voltages to the load when switched on. In this case using GE C137 SCR solving for R (firing the SCR of 12V of the 650V peak). I $gate = 40_{ma}$ for C137.

$$R = \frac{E}{I} = \frac{12}{40\,ma} = .3K \text{ or } 300\,\Omega$$

$$\text{peak power} = \frac{E^2}{R} = \frac{650^2}{300}$$

$$= 1{,}410 \text{ watts !!}$$

This is an impractical dissipation.

C34 - The LM301 has a short circuit limitation of 20 ma. If the gate is specified to be .2ms wide and the supply if $\pm 12V$.

then:

$$\text{across C34 } \frac{dv}{dt} = \frac{24V}{.2ms}$$

$$\text{now } i = C\,\frac{dv}{dt}$$

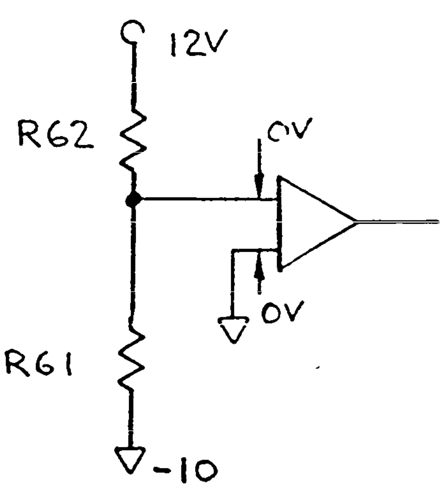
$$C = 20\text{ ma.}\ \frac{.2ms}{24}$$

$$C = .167\mu f \text{ use } .2\mu f$$

R61 . R62

what is desired for SCR to fire at 12 volts

is:



$$I = \frac{12}{R62} = \frac{10}{R61}$$

now R62 is arbitrarily picked for 470K since $\frac{440V}{470K} = 1ma$

so:

$$P_{R62} = 1/2 \text{ watt}$$

$$\frac{18}{470K} = \frac{12}{R61} \qquad R61 = \frac{12}{8}\ .470K$$

$$R61 = 705K \text{ use } 680K\,\Omega$$

D) Design Calcullations (Continued)

TR6

Since 40 ma are needed to fire C137 SCR's, then the 20 ma available from the LN1301 is insufficient. TR6 is a current step up. For 100% safety factor, I gate (pulsed) should be 80 ma.

$$\frac{NP}{Ns} = \frac{Is}{Ip} = \frac{80ma}{20ma} = \frac{4}{1}$$

or a 4:1 turns ratio.

This ratio also reduces the voltage from 24 to 6. Other components ar chosen by selection.

Transistor Switch

Refer to Drawing No. 1002, Figure IV-C.

The *main* function of the transistor switch is switch the current 30 amp at 650 volts. Q10 through Q19 are DTS804 (Delco) transistors; they have a maximum collector current of 5a and voltage of 800v (SUS). Th are configured as a double darlington since at 3 amps the DC B is about 4. One driver then dirves a bank of four with 3 amps base current de- livering 12 amps total collector current (3 amps per transistor). Ther fore, all transistors, *including* the drivers Q10 and Q15, have equal loads. R49 through R57 are emitter resistors for current balancing. Since the switch drives reactive loads (SR1, SR2 and TR3), D32 is used to conduct reverse *currents. D31* prevents the transistors from being operated in *a* reverse mode when current is flowing through D32. D29 and D30 are used to sweep out stored charge in the transistor banks and D28 sweeps *stored charge out* of the drivers.

As was mentioned in the previous discussion on the main assembly (Figure IV-A), a time delay between the "turn-off" of switch one and "turn-on" of switch, two, and vice versa, is needed to allow for capac- itor recovery of C5 (and C6) by reactor SR1. This display is pro- vided by the circuitry preceding the transistors (Figure IV-C) SCR4, Q9, C30 and related components. The input switch drive is a 6 volt peak symmetrical square wave. The transistors Q10 and Q15 received drive through SCR4 and R-46; however, SCR4 does not conduct until it receives gate drive via Q9 and R43. With the positive edge of the drive signal C28 is charged to *-6V* (previous level of drive square wave). R4 and C28 provide a fixed 4 sec delay in drive to Q9 (and thus the tran- sistors Q10, R15). During this time, C5, C6 (Figure 4A) begin their recovery of discharge. Their recovery time is variable, depending on output load; the worst recovery is 30 sec. (see discussion on main assembly, Figure IV-A). C30 senses the "negative sloping voltage (capacitor recovery) and feeds a negative current through R45, causing a negative voltage on the base of Q9, thus holding Q9 off. When the capacitor (C5) is fully recovered, the voltage on the anode of D31 cease its negative rate of change of voltage, current stops through C30 and Q9 receives drive through R11 and R45. SCR4 drives the QIO and Q15, driving the whole switch on.

D) Design Calculations

Emitter Resistors R49 - R57

Since VBE on DTS804 can vary from .65 to .75 volts, then equalizing resistors will be needed to equalize base and collector currents. Full load collector current is 3 amps per transistor at B = 4 @Ic = 3a, then IB = .75a and IE = 3.75a and allowing a 5% difference in emitter currents between the maximum spread of VBE. Since bases are pushed together, the base to ground voltage is the same for paralleled transistor (Q11-Q14, Q16-Q19).

For example:

$$V_B 11 = V_B 12$$

$$VB11 - IE_{11} R_{49} + VBE_{11} = IE_{12} R_{50} + VBE_{12}$$

if VBE11 = .65 and VBE12 = .75 and R49 = R50 and IE12 = 3.75-5%.3.7

then

$$3.75.RE + .65 = 3.56.RE + .75$$

$$.19\ RE = .10V$$

$$RE = .53\ r$$

use .5r resistors

R46 Drive Pass Resistor

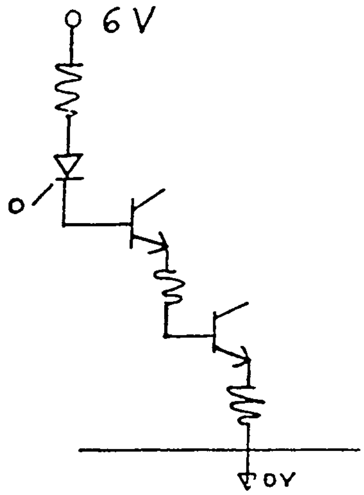
Now worst case B = 4 at IC = 3a in a darlington 4.4 = 16 I max = 3a/tra sistor or 30a total

$$IB\ total = \frac{IC\ total}{B\ total} = \frac{30}{16}$$

$$IB = 1.84\ amps$$

D) <u>Design Calculations</u> (Continued)

equivalent ckt



with 6V peak square wave and SCR4 on and IC = 30 then
VR46 = 6 - (circuit drop)

VSCR4 "on" = .95
VBE Q10 = .70
VR48 = 3a .2r = .6
VBE Q11 = .70
VR49 = 3.75(.5) = 1.87

circuit drop = 4.72

$R46 = (6 - 4.72)/1.87 = \frac{1.28}{1.87} = .695$

use <u>.8Ω</u>

<u>R44, C28</u>

C28 must change from -6 through +5 in approximately 4 seconds for a 4 sec. fixed delay.

$\tau = R44\ C28$

$VC = V\max\left(1 - \exp\left(-t/\tau\right)\right)$

$11 = 12\left(1 - \exp\left(-4/\tau\right)\right)$

$12 \exp(-4/\tau) = 1, \ \exp(-4/\tau) = 1$

$-4/\tau = \ln\frac{1}{12}$

$-4/\tau = 2.49$

$\tau = 1.6$ sec.

Pick $C_{28} = .01\mu f$

$\tau = R44\ (28 = 1.6\ \mu\ sec$

$R44 = 160\ \Omega$

use $C_{28} = .01\mu f \quad R44 = 150\ \Omega$

D) Design Calculations (Continued)

C30 Recovery Sense

If C5 recovers in 30μsec (650V)

then

$$\frac{dv}{dt} = \frac{650}{30\mu sec} = 21.6V/$$

to hold Q9 off during recovery IC30 must reverse bias the base of Q9

$$Ic30 = Ir45 \qquad \text{now } Vr45 \approx 8V$$

to cut off Q9

$$Ir45 = \frac{8V}{4.3K} = 1.86\,ma$$

$$L = C\frac{dv}{dt}$$

$$C_{30} = (1.86ma)\frac{\mu sec}{21.6V} = 86pf$$

Use C30 = 120 pf

Logic Power Supply

Refer to Dralwing No. 1003, Figure IV-D.

The logic power supply's function is to supply the voltage at required currents to the switch drivers (Figure IV-E), the timer module (Figure IV-F, and the Feedback Amplifier circuit (Figure IV-G). The operation is straight forward. The regulators receive power from TR1 9shown on Figure IV-A) whose secondary is centertapped 10 VAC and 18 VAC secondary. The 18 VAC windings are full wave rectified by D10 and Dll; this unregulated line is filtered by C3 (shown on main assembly Figure IV-A) and supplies 24 volts to the switch drivers (Figure IV-E). R7 is a surge limiting resistor. The 24 volt line is rated at 4 amps. The 10 VAC windings are full wave rectified by D12 to D15, and filtered by C10 and C R9, R10, D16 and D17 are zener diode regulators which provide +8 volt lines for the feedback amplifier (Figure IV-G). The +8 volt lines are rated at 60 ma. IC1 is an LM 309K voltage regulator for the 5 volt supply for the TTL IC's in the timer module (Figure IV-F). The five volt line is rated at .5 amp.

D) Design Calculations (Continued)

24 Volt Supply

18 VAC RMS supply VDC = V peak = 1.41 x 18 V

V peak = 25.4 V

Open circuit voltage = 25.4 V
Full load voltage = 25.4 V - Vdiode - IR7
= 25.4 - .85 - .4
= 24.1 volt neglecting transformer losses

5 Volt Supply

10 VAC RMS supply VDC = V peak = 1.41 x 10 14.1V

V_{in} for LM309K - V_{in} = $V_p - 2V_b - .47\Omega I$
= 14.1 - 2(.75) - .24
= 12.4V

This is ample supply for the LM309K.

$\pm$ 8 Volt Supply

V peak = Vdc = $\pm$14.1V

R10 = R9 = Vr10/I max = $(V_p - V_b - V_z)/I\ max$
= (14.1 - .75 - .8) 160 ma
= 90Ω

R10 = R9 = 100Ω

Switch Drivers

Refer to Drawing #1004, Figure IV-E.

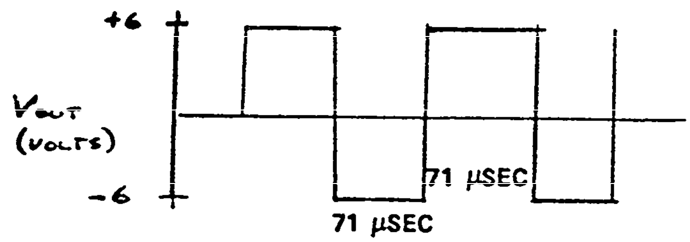
The operation of this circuit is straight forward. The drive signal to the switches is transformer coupled to the transistor switches through TR4. There are two secondary windings, one winding for each transistor switch in a complementary pair. The 24 volt supply line comes in through pin Z; it is decoupled by C12, and continues to the primary centertap of TR4. TR4 is a centertapped driven inverter transformer, with the main drivers being Q2 and Q4; R15 is a common emitter resistor. The low level drive signals come in 180° out of phase of each other at pins 1 and 15. Q1 and Q3 are connected to Q2 and Q4 respectively in a darlington configuration. R11, R12, R13 and R14 are pull down resistors for the transistors.

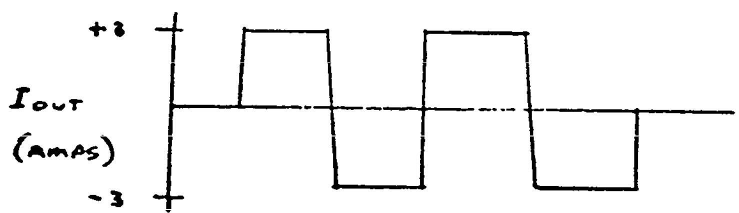
D) Design Calculations (Continued)

Switch Drivers (Continued)

TR4

The switch driver must supply the following waveform to the transistor switches:





The drivers must supply 6.0 volts at 3.0 amps max to the transistor switch. This is 18 watts per winding or 36 watts total.

Core

Area = $.6CM^2$ B max = 5600 gauss
V applied = 24V

Primary turns

$$10^8 V = 4 \text{ NpA B max f}$$

$$Np = \frac{10^8 \cdot 24}{4 \cdot .6 \cdot 5600 \cdot 7000} =$$

$$Np = 25.51 \text{ turns}$$

For safety use Np = 36 turns bifilar

D) <u>Design Calculations</u> (Continued)

Secondary Turns

$$\frac{N\ sec}{N\ pri} = \frac{V\ sec}{V\ pri} \qquad N\ sec = 36 \bullet \frac{6}{24}$$

$$N\ sec = 9\ turns$$

<u>Q2, Q4</u>

$$Ip\ max = \frac{Np}{Ns}\ I\ sec = \frac{9}{36}\ 3 = .75\ amp$$

Therefore, Ic max = .75a. This is within Ic max of 2N3879.

<u>Ql, Q3</u>

$$Ib_{min_{2,4}} = \frac{Ic\ max\ 2,4}{B\ min_{2,4}} = \frac{.75a}{20} = 37.5\ ma$$

$$Ib_{min_{2,4}} = Ic\ max_{1,3} = 37.5\ ma$$

This is within Ic max of 2N3567.

<u>Timer Module</u>

Refer to Drawing #1005, Figure IV-F.

The basic function of the timer module is to 1) set up 7000 Hz clock frequency of the unit; 2) receive an error signal from the feedback circuit; 3) adjust the phasing of the transistor switches (see Figure III-W and related discussion).

R16, C13 and IC2 make up an astable flip flop operating at 14 KHz. C15, C16, C17 and IC3 (1) and (2) perform waveshaping functions. R21 and C18 integrate the error signal and feed it to IC4, which is a voltage controlled time delay. The fixed clock and the delayed clock IC2, and IC4 respectively operate JK flip flops (IC5). The JK flip flops drive emitter followers Q5 to Q8, which supply the low level signal to the switch drivers. IC3 (3) and (4), and C19 perform a phase locking function to insure that Q6 always lags Q8. The operation can be more easily seen in a waveform progression shown in Figure IV-P. Waveform VC13 shows Cl 3 oscillating in a relaxation mode. VC16 shows C16 moderately degrading the square edges of the output of IC2. IC3 (2) shows a narrow one-shot pulse; this pulse is used to reset IC4 if necessary. Vc20 shows the voltage across C20; C20 charges through R22 to the voltage (control feedback voltage or integrated error signal) at pin 5 at which time it discharges, creating a time delay Ø which phases the signals to the switch drivers. If for some reason Vc20 ncver reaches V pin 5 (very large error signal, for example) then the reset pulse 1c3 (2) resets the timer. thus limiting a maximum phase separation of 180.

D) Design Calculations (Continued)

R17 R16 C13

$$f = \frac{1.44}{(R17 + 2R16)C} \qquad \text{based on 555 characteristics}$$

let C = .1μf

$$R17 + 2R16 = \frac{1.44}{(.1 \cdot 10^{-6})(14.10^{3})}$$

$R17 + 2R16 \approx 1\ K\Omega$

Pick R17 = 560

$$R16 = \frac{1K - 560\Omega}{2} = 220\ \Omega$$

$R16 = 220\Omega \quad R17 = 560\Omega$

C15 R18

These components determine shape of the IC3 (2) waveform (pulse see Figure IV-P).

Pulse width = .9 τ where τ = (R18)(C15)
if pulse width = 10μsec

$$10 \cdot 10^{-6} = .9 \cdot R18 \cdot C15$$

let C15 = .0015μf

R18 $\cong$ 7.4K use 8.2 KΩ

C15 = .0015μf R = 8.2Ω

C16 R20

These components must delay the set pulse (Vc16, Figure IV-P) tc pin 2 until the 555 timer is reset by IC3 (2), about 4 sec.

Time delay $\cong 2\tau$ where τ = (C16) (R20), now R20 must be less than 150 in order to insure logic level transmission from Ic2 Pin 3 to Ic4 Pin 2.

Let R20 = 82Ω

$$\tau = \frac{4\mu sec}{2} = 2\mu sec = C16\ (82)$$

$$C16 = \frac{2\mu sec}{82} = .024$$

use C16 = .02μf

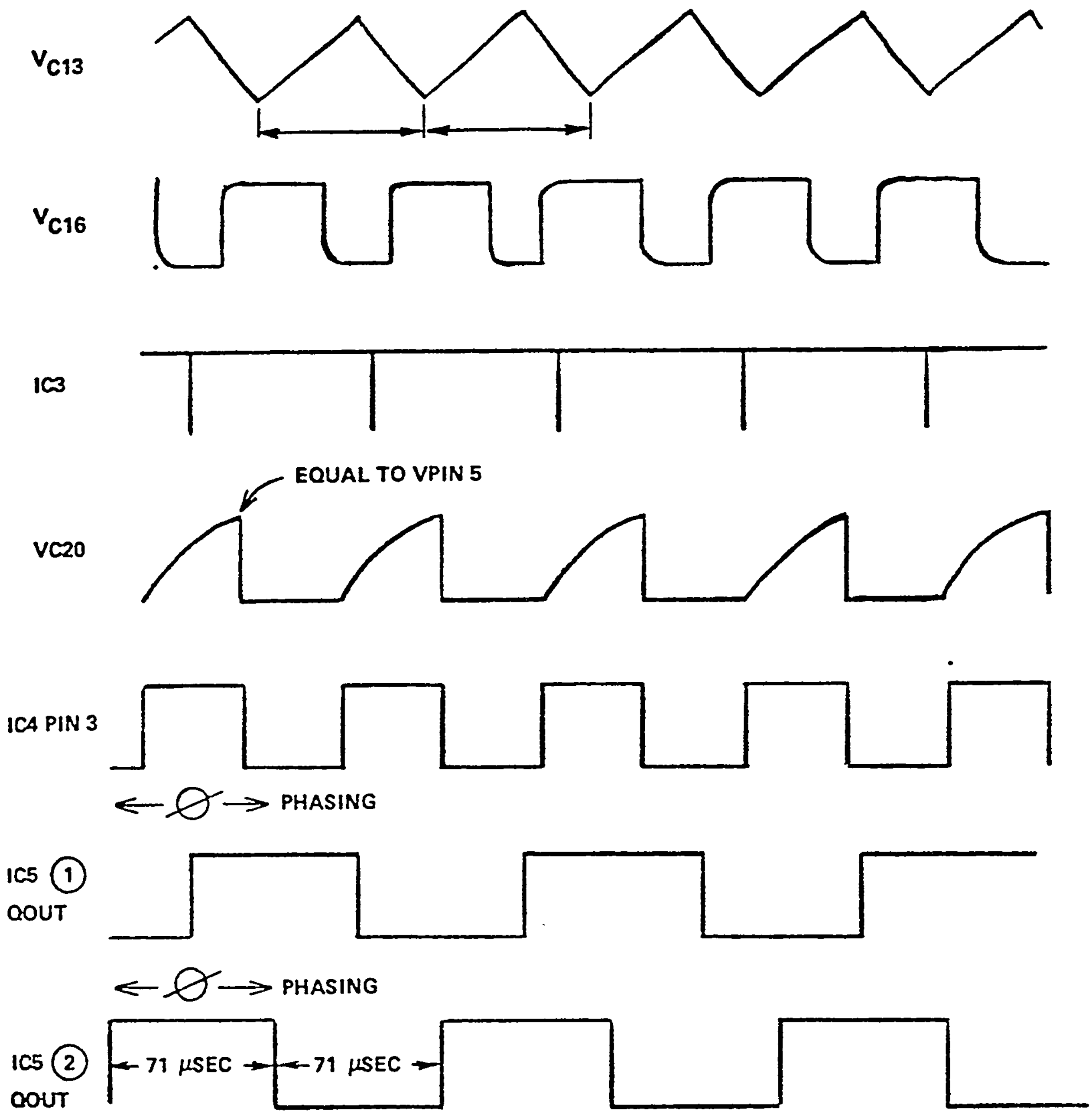

V_{C13}
V_{C16}
IC3
EQUAL TO VPIN 5
VC20
IC4 PIN 3
PHASING
IC5 1
QOUT
PHASING
IC5 2
QOUT
71 µSEC
71 µSEC

Figure IV-P

R22 C20

These components determine delay time Ø of Ic4 pin 3, Figure IV -P.

Since V pin 5 = 5 volt, C20 should charge to 5 in 71μsec.

charge time $\approx 2\tau$ where τ = (R22) (C20) pick C20 = .001μf.

$$.71\mu sec = (R22 \cdot .001\mu f)2$$

$$R22 = 35.5K$$

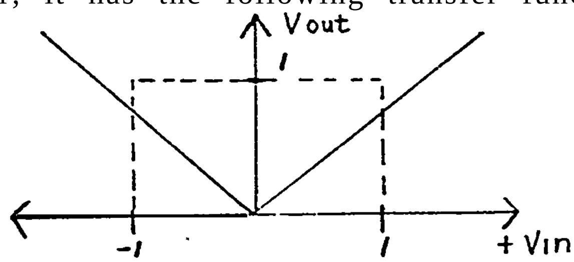
use $R22 = 36K$

Other components are pull-up resistors or filtering capacitors.

Feedback Board

Refer to drawing #1006, Figure IV-G.

The basic function of this circuit is to convert a low level signal from a shunt, to a workable level for the timing module. Al is a lossless full wave rectifier; it has the following transfer function:



When the input (pin 4 to R28) is positive, the output of Al saturates negatively. D25 blocks the negative amplifier output and the signal passes through R2S and R32 to the output. If the signal in is negative the D25 is forward biased by the amplifier output and the circuit acts as a unity gain inverting amplifier, thus the output is again positive. A2 performs as the summing junction of the reference signal (pin 5) and the controlled signal (amps). A2 also performs a pure integration making this a type 1 system with zero (or approaching zero) static error. The output of A1 goes to the timing module. Since the output of the full wave rectifier (Al) is high impedance, the amplifier A3 is a high input impedance (relative to 2 out of Al circuitry) unit gain amplifier used to drive the meter.

IV E) Mechanical Design

1. Heat

An estimate was made of anticipated component heat dissipation for purposes of thermal analysis and heat sink selection. This estimate is as follows:

1.	Overload control circuitry	50
2.	Input bridge	200
3.	Transition switches	390
4.	Output bridge	780
5.	Transformer coil	80
6.	Control logic and switch driver	20
7.	Clipper Network	80
8.	Various other components	neg.
	TOTAL	1600

Since approximately 85% of the heat dissipated is in the output bridge, transistor switches and input bridge, they will be the primary considerations for a detailed thermal analysis. For purposes of this calculation, maximum ambient air was assumed to be 50 C (122° F). This would include the effects of maximum temperature shopair possible being heated be adjacent power supplied or related equipment.

a) Output Bridge

Preliminary calculations indicated that heat dissipation in the output bridge would be a problem. As a result the original D09 size stud mounted type rectifiers were replaced with GE A396 Press Pack type rectifiers. The press pack configuration allows for two sided mounting with the result being improved heat transfer characteristics. Thermal characteristics for this device are:

GE A396 Press Pack Rectifier	
Max Junction Temp	125C
Thermal Resistance J-C	15C/Watt

Two rectifiers dissipating 780 watts amounts to 390 watts per rectifier. Initial estimates of thermal resistance of a typical press pack type heat sink @ .18C/watt. Insulating washers will be required between the components and the heat sink. Because of the high heat transfer conventional insulating washers were bypassed for beryllium oxide type which have superior heat transfer characteristics. Using a .062 thick beryllium oxide washer the size of the contact area of the press pack rectifier and including contact resistance, a total thermal resistance of 12°C/watt per rectifier was calculated.

E) Mechanical Design (Continued)

a. Output Bridge (Continued)

A thermal rise calculation can then be made for each device.

max ambient temperature		
heat sink T.R	.18C/watt x 390 watt	= 50C
insulator T. R.	.12C/watt x 390 watt	= 70.2C
rectifier T.R.	.15C/watt x 390 watt	= 58.5C
Total junction temperature		225.5C

As can be quickly seen, this greatly exceeds the rated maximum. allowable junction temperature. To remedy this solution, it was decided to increase the number of rectifiers to four instead of two. By paralleling the devices, the heat dissipation per device would be cut *in* half, resulti in a calculated max junction temperature of 170C. This still exceeds the rated temperature. However, after further evaluation and testing of the breadboard, the power estimates were reduced to about 100 watts per device with four devices. This results in a maximum junction temperature of 113°C, which is below the maximum component rating allowed for this device.

A heat sink was then selected which would provide the required heat dissipation. This extrusion was Thermally 6740 or equivalent Wakefie 3560, which are designed specifically for press pack mounting.

Wakefield 3560
Weight 4.3 lb./ft.
T.R. = .18C/watt (based on a pair of extrusions 14" long - two or more devices per assembly to distribute heat. (See Wakefield Cat. No. 103, page 47).
Thermalloy No. 10888 compression clamps were used in conjunction with the heat sinks for proper mounting.

Transistor Switches

The switching circuits to be heat sinked consist of forty (40) TO3 case transistors. These transistors conveniently fall into four groups of 10 each. A quick layout indicates that 10 transistors arranged closely in one row will fit within a 14" length with some room on the end for mounting holes. Calculations are based on mounting one row of 10 transistors along with 2-D04 diodes per heat sink. Heat dissipation is based on 13 watts per transistor.

Using a Thermalloy 6175 extrusion, a Thermal resistance of .56°C/wat was calculated, based on distributed components.

The thermal rise calculation then becomes:

Ambient air temp	50C
Heat sink rise .50°C/W x 130W	73C
Kapton insulators .5°C/W x 13W	7C
Case to junct rise 1°C/W x 13W	13C
Total max Junct temp	143C

E) Mechanical Dcsign (Continued)

Transistor Switches (Continued)

This is well below the maximum allowable junction temperature of 200°C for the Delco DTS 804 transistors. Associated components mounted to the heat sink will contribute very little heat.

Input Bridge

The input bridge consists of a set of three each syntron R34100 DO5 case size diodes and a set of syntron S34100 DO5 case six diodes. Power dissipation is 200 watts total or 34 watts per device. Thermal resistance is 1°C/watt and maximum junction temperature is 2000C.

Since there is less heat dissipated by three of these diodes than a row of the 10 transistor switches, it was decided to use the same extrusion (Thermalloy 6175) for uniformity in packaging and mount each set of diodes on a separate heat sink

2. Packaging

With the heat sinks selected and a basic concept in mind, a package layout was made. Heat sinks were arranged in a compact configuration with the height being determined by the transistor switch configuration as 14 inches. Heat sink layout is shown in Figure IV An external tubular frame was designed around the heat sinks as the main support. Overall height of the unit of 17 inches al for 1-1/2 inches of free air space below and above the heat sinks for improved air flow. Final package dimensions for the breadboa were: 15.5 inches wide x 21 inches long x 17 inches high. The internal case allows pleanty of room for the remaining components The power transformer being the next consideration was mounted flat on the bottom of the inner case (See Figure IV-Q) and bolted through the case floor directly to a frame cross piece for rigid support. This helped to keep the center of gravity low and central located. The frame structure also provided convenient carrying handles on each side of the unit. By bussing the output rectifier current to the lower end of the heat sink, all of the heavy gage wiring could conveniently be located at the bottom of the inner case then be routed to the precision shunt and output connectors on the lower front panel. (See Figure IV-Q). A phenolic platform was then situated above the transformer to provide mounting for the remaining electronics. These electronics would be located during assembly.

The final package showing the front panel is illustrated in Figure IV-R.

3. Weight

A preliminary weight analysis was conducted using the projected design for the breadboard package. These weights were estimated as follows:

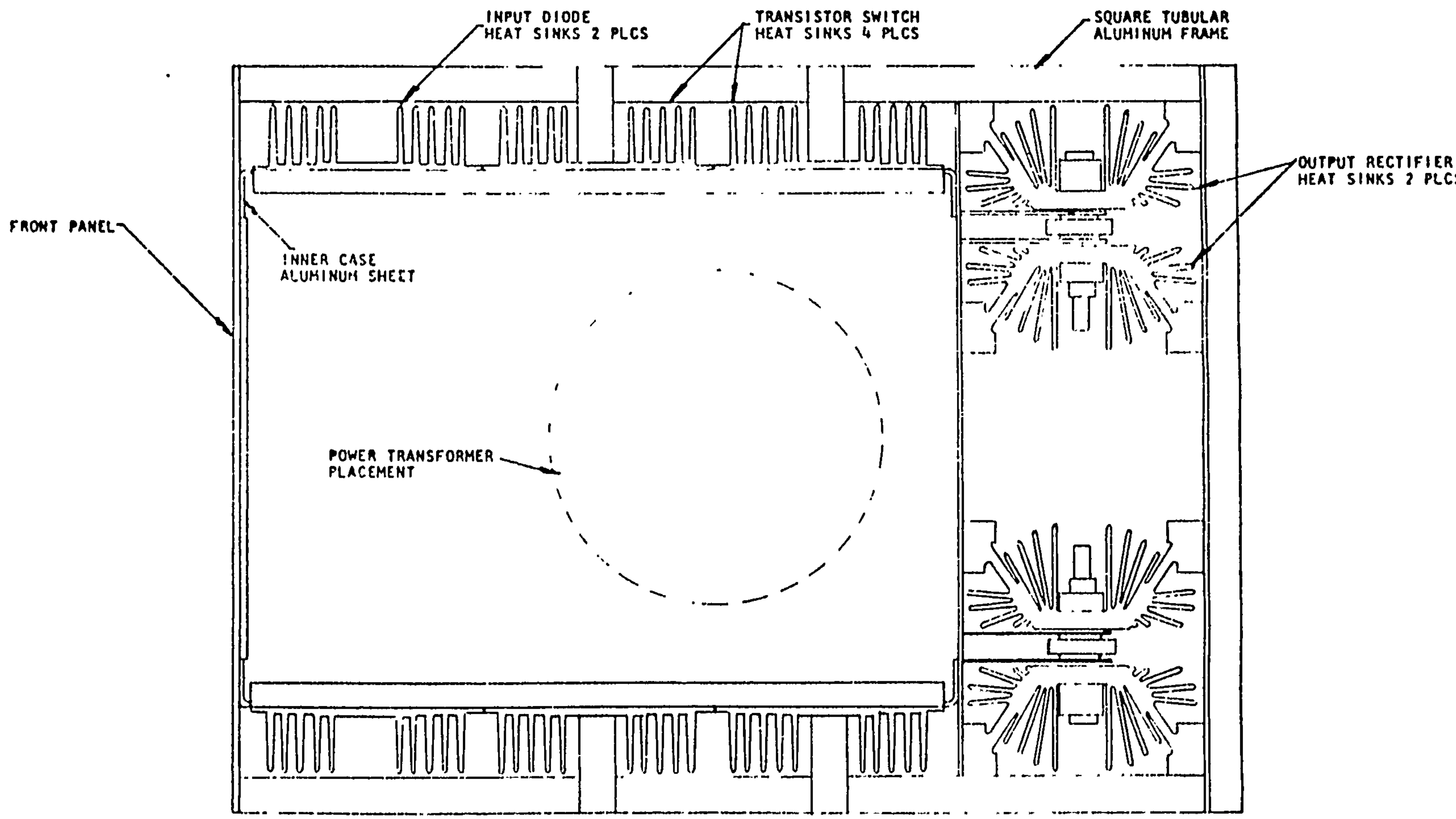


Figure IV-Pa
Heat Sink Configuration

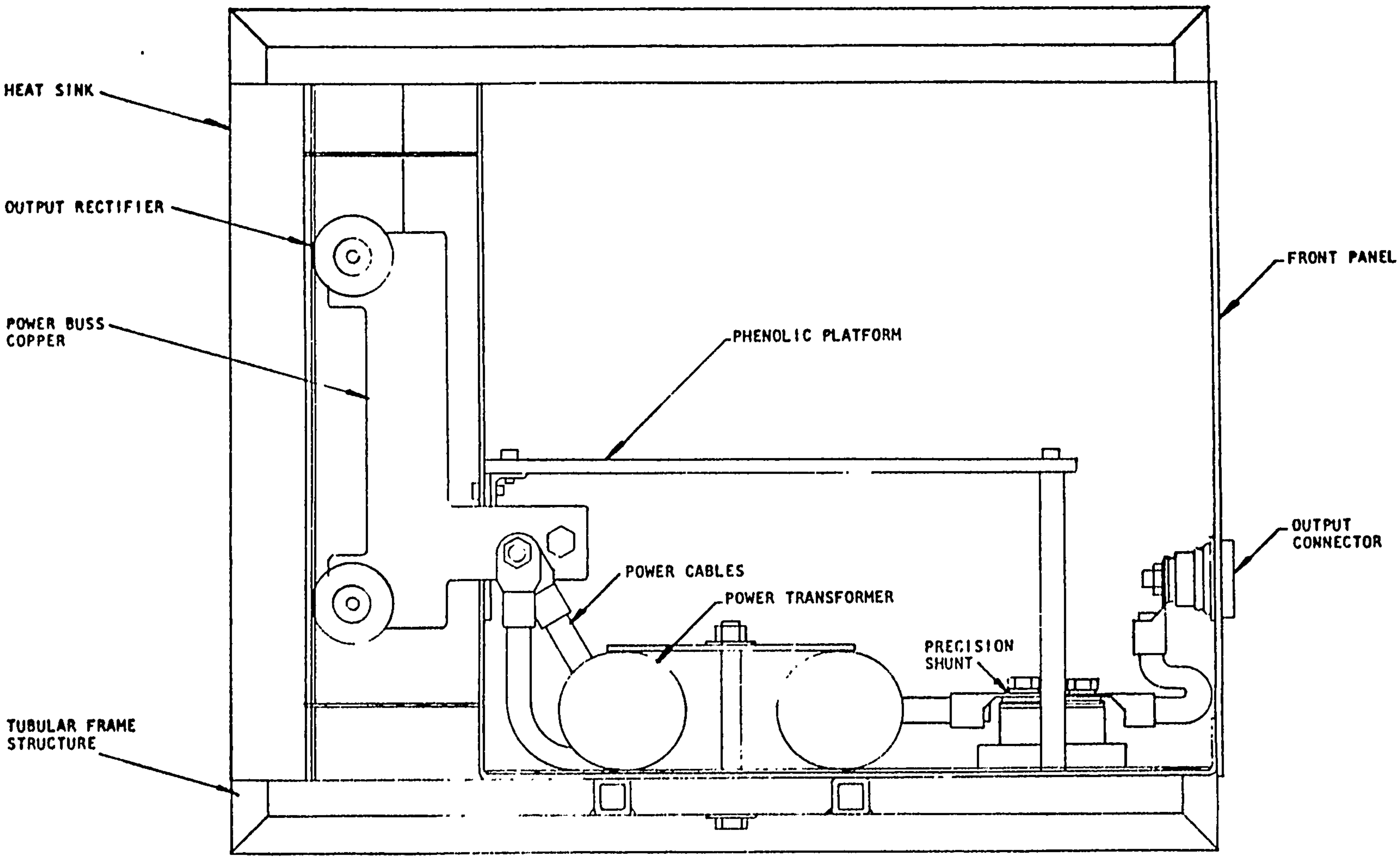


Figure IV-Q
Power Transformer Placement

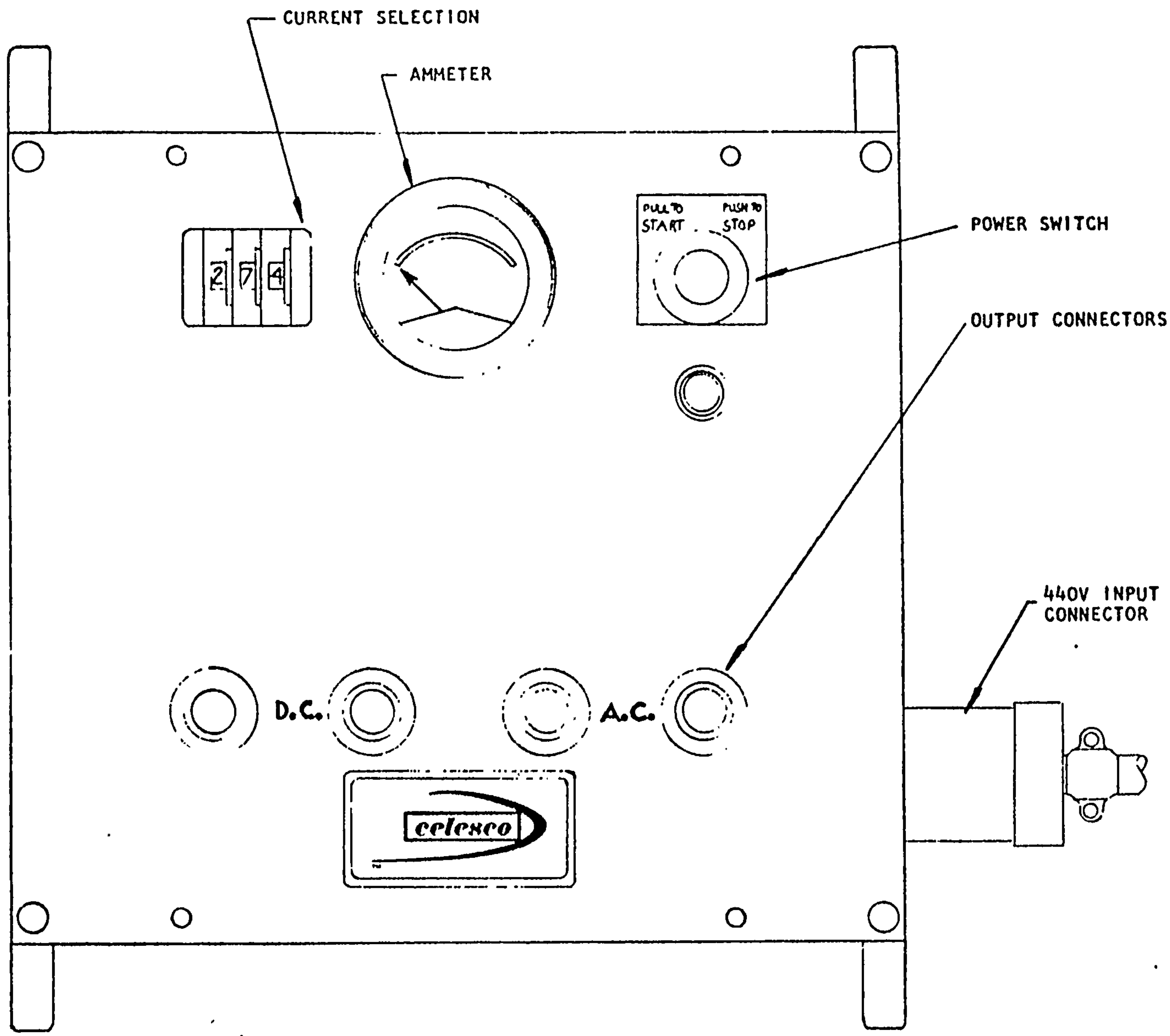


Figure IV-R
Front Panel Configuration

E) Mechanical Design (Continued)

3. Weight (Continued)

WEIGHT ESTIMATE

Frame and Structure	13.4
Front Panel	2.7
Heat Sink Extrusions	35.2
Power Transformer	7.5
Electronics	14.9
Wire and Buss	3.6
Switches, Connector, Meter	2.4
Hardware and Insulations	6.4
Total	86.1 lbs.

F) Final Package Performance Test Results

The final package was tested for resistive loads, as well as under welding operations. Temperature tests were made under both conditions. Resistive (load bank) operation was at full output for twenty minutes continuous. Weld operation was seven rods of 6011 3/16" rod (about 20 minutes). The temperature rises are listed below:

	Welding	Load Bank
Transistor banks	25°C	20°C
Output Rectifiers	20°C	30°C

Temperature rise of heat sink from ambient temperature.

As can be seen, the welding operation is more stressful to the transistors, due to welding transients (as evidenced by increase in temperature rise).

The welding performance of the machine was tested. The unit could successfully perform the following welding processes:

1. Stick (downhand) up to 3/16" rod
   a. Straight polarity DC
   b. Reverse polarity DC

2. T16 welding 40-250 amps (argon cover gas)
   a. Straight polarity DC
   b. Reverse polarity DC

3. Mig welding - .032 aluminum wire, helium cover gas

Note: There was no slope or inductance control incorporated into the control circuits.

F) Final Package Performance Test Results (Continued)

This section of the report corresponds to points four and five (a) of the Plan outlined in the objectives of this report.

V Conclusion

A) Evaluating of Final Package

The conclusion of the results of a project such as this must ultimately relate to the objectives set down at the onset of a project. Not all of the objectives were proven explicitly, due to the limited time remainin for testing. However, results from the tests that were made can be extrapolated into other areas.

1. Welding Performance

The basic module could supply 100a AC at 7000 Hz. This falls short of the 250 amp AC goal. Test indicated, however, that 7000 Hz welding may be difficult, and have adverse effects on the weld and welder. However, AC welding at 60 Hz may be established from 7000 Hz using electronic techniques, thus eliminating transmission and welding difficulties.

The basic module could supply 250a DC and have the capability of being paralleled with other modules to provide up to 1000a DC. Only one unit was fabricated, thus parallel capability could not be established explicitly; however, the nature of the output circuitry is such that parallel capability (DC) may be conducted, though it is not proven explicitly.

Welding operations of downhand stick Mig and Tig welding were performed. Due to lack of time, other processes mentioned in the objectives were not tested explicitly, but there is reason to assume performance in the other categories.

From a welding standpoint then, the final unit did fulfill a portion of the desired objectives.

2. Electronic arid Maintainability Performance

The unit operated from 440 VAC 60 Hz line power and was capable of compensating for fluctuations in primary line voltage. A closed loop feedback system was utilized for this function.

Visual failure signals were not incorporated into the final unit nor were remote control capabilities. These functions are, from a design point of view, relatively easy to incorporate into a system

V A) 2. Electronic and Maintainability Performance (Continued)

The basic module final weight was 98 lbs. and rectangularly shaped for easy handling. Duty cycle was less than 100%. The basic module was free from vents and openings and cooled by convection only. Water cooling and forced air cooling were not used.

Power supply components consisted of plug-in modules in the control system portion. The final unit was of prototype nature and thus not capable of withstanding rough handling. This capability, however, *can* be incorporated into production units. There is nothing inherent in "electronic" components that make them necessarily fragile.

From an economic and reliability point of view, the electronic design utilizing transistors is not the optimum design. It is felt that SCR's provide a cheaper, more reliable means of switching than transistors, providing they *can* be efficiently controlled at the frequencies needed for this design concept.

B) Recommendations for Future Study

Since the unit that the Celesco Engineering staff produced as a result of a project initiated by the Marad Commission demonstrates feasibility more so than producibility, further work is necessary to reduce the now proven concept to practice. The following recommendations for future study can be made:

1. Reduce the input line voltage to 220V. The 440 volt line pushed state-of-the-art semiconductor ratings. By reducing the input requirements by half, reliability is considerably improved. Since the concept affords high efficiency at power factors close to 1, a 300 amp lighfweight supply would draw about the same current from a 220V, 3 phase line. Thus the voltage on power cables could be reduced, thus increasing safety without increasing wire size.

2. As mentioned, SCR's should be investigated in the light of what has been learned by the transistor investigation.

3. SCR's and electronic techniques should be investigated to convert high frequency of conversion back to 60 Hz for AC welding.

4. An interest has been expressed in a low current (to 150a) 120 VAC 60 Hz single phase "mini" supply. This should also be investigated.